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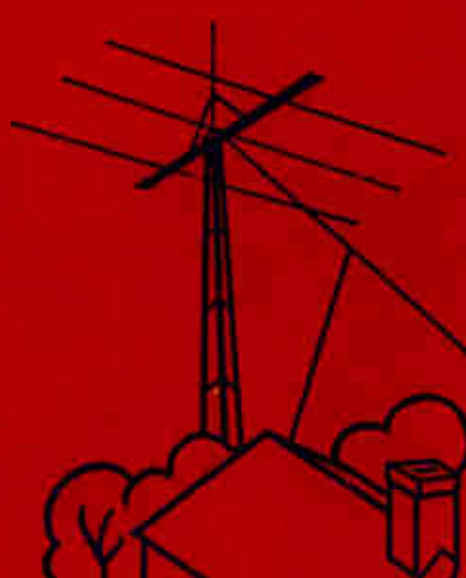
1.00

The radio amateur's handbook

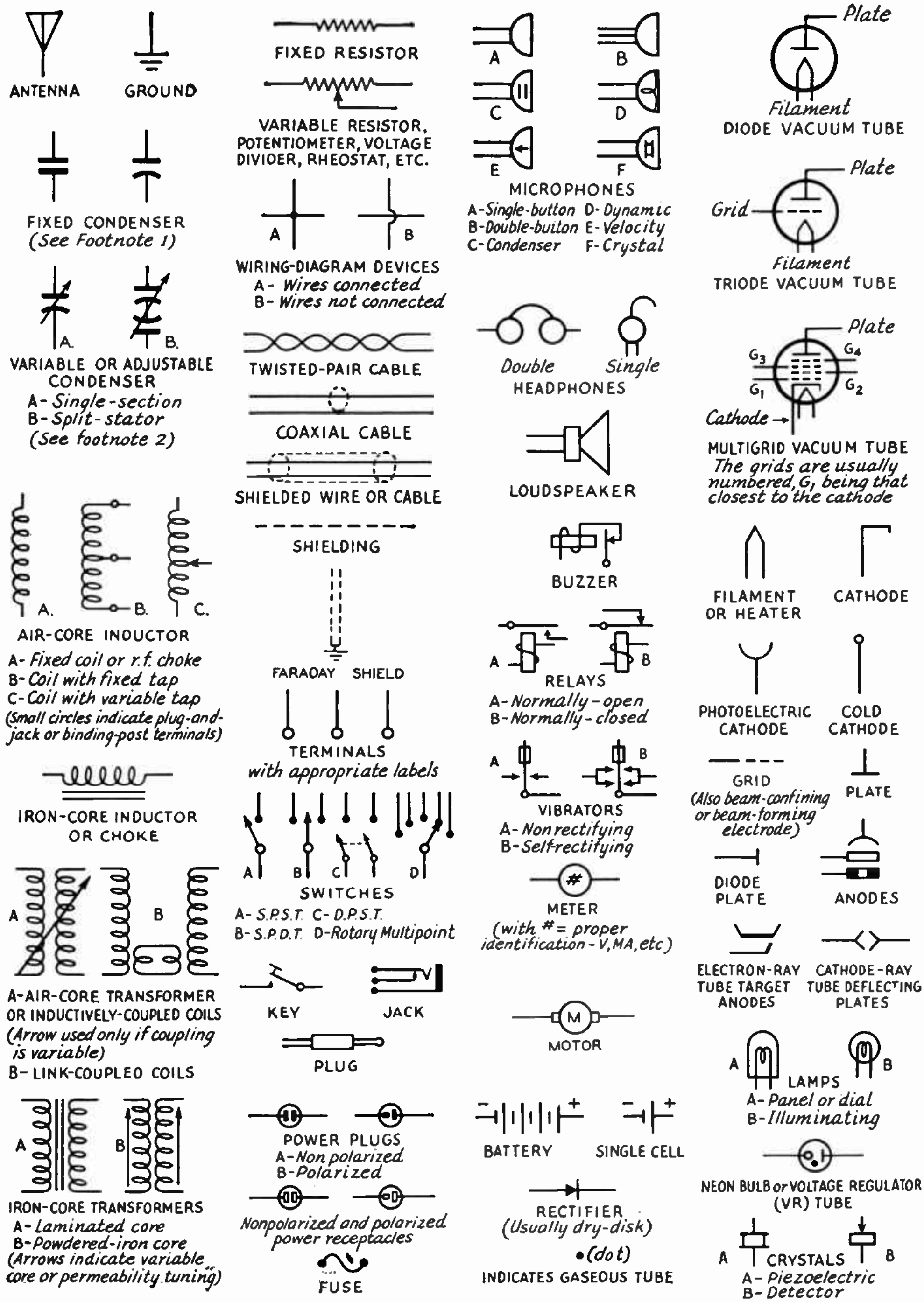
THE STANDARD MANUAL OF AMATEUR RADIO COMMUNICATION

$2.00



PUBLISHED BY THE AMERICAN RADIO RELAY LEAGUE

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.

The radio amateur's handbook

By the

HEADQUARTERS STAFF

of the

American Radio Relay League

West Hartford, Connecticut, U.S.A.

1950

Twenty-Seventh Edition

Twenty-Seventh Edition

THE RUMFORD PRESS
CONCORD, NEW HAMPSHIRE, U. S. A.

Foreword

This twenty-seventh edition of *The Radio Amateur's Handbook* is the latest of a series extending over twenty-four years of continuous publication, a period during which the total circulation has climbed to well over two million. The immediate and enthusiastic acceptance of the first edition by the radio amateurs of 1926 has been matched by continuing popularity throughout the intervening years — a popularity based on the *Handbook's* practical utility, its treatment of radio communication problems in terms of how-to-do-it, and its long-established policy of presenting the soundest and best aspects of current amateur practice rather than merely the new and novel. These same features have won for the *Handbook* universal acceptance in other segments of the technical radio world — engineering, educating, servicing, operating — even though the book is written primarily for the radio amateur. Its preparation and production is the work of the headquarters staff of the amateur's own organization, the American Radio Relay League.

The current edition reflects the changes that have taken place in the technical practices of amateur radio during the past year. Of major concern to amateurs in practically all the larger centers of population is the problem of interference with television reception, a subject that is treated extensively in this edition. Equipment that is designed to be as harmonic-free as possible is featured in the chapter on construction of transmitters, and new material on harmonic reduction has been included in the antenna chapter. The growing importance of single-sideband telephony has resulted in an increase in the space devoted to this subject. The chapter on measuring equipment has been expanded, in line with the widespread interest in — and necessity for — reliable measurements at both low and high frequencies. A considerable amount of new equipment is incorporated in the chapters covering the very-high and ultra-high frequencies. And as always, the tube tables have been revised to incorporate the new tubes that have appeared during the year.

Those to whom the *Handbook* has for years been an indispensable companion are well aware of it, but for new readers it is worth pointing out that in contrast to most publications of a comparable nature, the *Handbook* is printed in the convenient format of the League's monthly magazine, *QST*. This, together with extensive and usefully-appropriate catalog advertising by reputable manufacturers producing equipment for radio amateurs, 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.

It is sincerely hoped that this new edition will succeed in bringing as much assistance and inspiration to amateurs and newcomers to the hobby as have its many predecessors.

A. L. Budlong
Secretary, A.R.R.L.

West Hartford, Conn.
December, 1949

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.

— *Paul M. Segal*

CHAPTER 1

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.

● TRANS-ATLANTICS

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 seven 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 League's seventy-two 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	dididididit	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	dahdididididit
2	dididahdahdah	7	dahdahdididit
3	didididahdah	8	dahdahdahdidit
4	dididididah	9	dahdahdahdahdit
5	dididididit	0	dahdahdahdahdah

Period: didahdidahdidah. Comma: dahdahdididahdah. Question mark: dididahdahdidit. Error: didididididididit. Double dash: dahdidididah. 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).

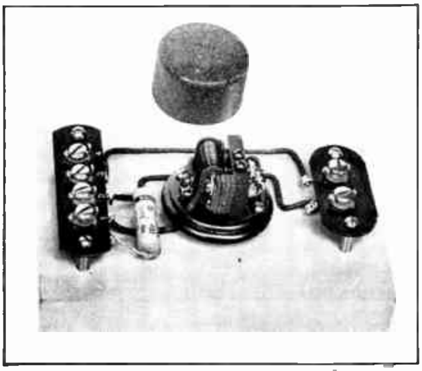


Fig. 1-3 — The cover of the buzzer unit has been removed in this view of the buzzer code-practice set.

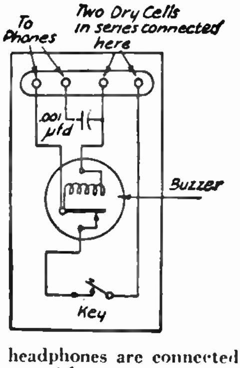


Fig. 1-2 — The headphones are connected across the coils of the buzzer, with a condenser in series. If the value shown gives an excessively loud signal, it may be reduced to 470 μμfd. or 220 μμfd.

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.

LEARNING BY LISTENING

As soon as you can, listen on a real communications receiver (with beat oscillator) and have the fun of learning by listening. W1AW conducts practice transmissions Mondays, Wednesdays and Fridays, 9 to 35 w.p.m., and Tuesdays–Thursdays, 15 to 35 w.p.m., starting at 9:30 P.M. EST. 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*. Practise 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.

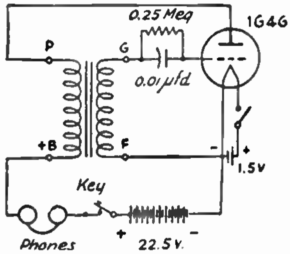


Fig. 1-4 — Wiring diagram of a simple vacuum-tube audio-frequency oscillator for use as a code-practice set.

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.

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 auto-

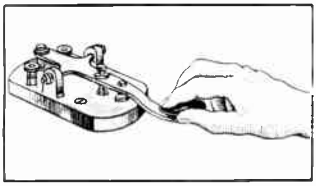


Fig. 1-6 — This sketch illustrates the correct position of the hand and fingers for good sending with a telegraph key.

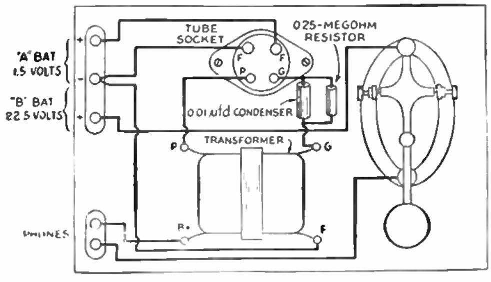


Fig. 1-5 — Layout of the audio-oscillator code-practice set. All parts may be mounted on a wooden baseboard, approximately 5 by 7 inches in size.

matic 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.

In the adjoining table 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, NFM designates narrow-band frequency- or phase-modulated radiotelephony, and FM means frequency modulation, 'phone (including NFM) or telegraphy. In addition, amateurs are assigned portions of the band 1800–2000 kc., subject to certain power and geographical restrictions, as shown in the table below.

3.500– 4.000	— A1
3.850– 4.000	— A3, Class A only
3.850– 3.900	— NFM, 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	— NFM, Class A only
26.960–27.230	— AØ, A1, A2, A3, A4, FM
28.000–29.700	— A1
28.500–29.700	— A3
28.500–29.000	— NFM
29.000–29.700	— FM
50.0 –54.0	— A1, A2, A3, A4
51.0 –52.5	— NFM
52.5 –54.0	— FM
144 –148	— AØ, A1, A2, A3, A4, FM
220 –225	— AØ, A1, A2, A3, A4, FM
420* –450*	— AØ, A1, A2, A3, A4, A5, FM
1,215 – 1,300	— AØ, A1, A2, A3, A4, A5, FM
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 1947 International Radio Conference resulted in certain planned changes in present bands which may become effective some time in 1950. They are: a reduction in the 20-meter band to make it thenceforth 14,000–14,350 kc., and a new band 21,000–21,450 kc. Further, in late 1949 there appeared to be substantial agreement between FCC and the amateur body on the desirability of certain additional changes in domestic regulations, among which are the extension of the 75-meter Class A 'phone band to read 3800–4000 kc. and extension of NFM privileges throughout the 50-Mc. band. Because of the possibility of such changes each amateur should keep himself currently informed by consulting *QST* or by writing ARRL for latest information.

(1) 1800 to 2000 and 2006 to 2050 kc. Use of this band by amateur radio stations is restricted as follows:

(i) 1800 to 2000 kc. Use of this band is on a shared basis with the Loran system of radio navigation. In any particular area the Loran system of radio navigation operates either on 1850 or 1950 kc., the band occupied being 1800–1900 or 1900–2000 kc. The amateur service may use in any area whichever bands, 1800–1825 and 1875–1900 kc., or 1900–1925 and 1975–2000 kc., are not required for Loran in that area, in accordance with the following limitations and conditions:

(*a*) Mississippi River to East Coast U. S. (except Florida and states bordering Gulf of Mexico): 1800 to 1825 kc. and 1875 to 1900 kc., using type A-1 or A-3 emission. Power input to the plate circuit of the tube or tubes supplying power to the antenna shall not exceed 500 watts day, 200 watts night.

(*b*) Mississippi River to West Coast U. S. (except states bordering Gulf of Mexico): 1900 to 1925 kc. and 1975 to 2000 kc., using type A-1 or A-3 emission. Power input to the plate circuit of the tube or tubes supplying power to the antenna shall not exceed 500 watts day, 200 watts night, except in the State of Washington where daytime power is limited to 200 watts and night-time power to 50 watts.

(*c*) Florida and states bordering Gulf of Mexico: 1800 to 1825 kc. and 1875 to 1900 kc., using type A-1 or A-3 emission. Power input to the plate circuit of the tube or tubes supplying power to the antenna shall not exceed 200 watts day, no operation at night.

(*d*) Puerto Rico and Virgin Islands 1900 to 1925 kc. and 1975 to 2000 kc., using type A-1 or A-3 emission. Power input to the plate circuit of the tube or tubes supplying power to the antenna shall not exceed 500 watts day, 50 watts night.

(*e*) Hawaiian Islands: 1900 to 1925 kc., and 1975 to 2000 kc., using type A-1 or A-3 emission. Power input to the plate circuit of the tube or tubes supplying power to the antenna shall not exceed 500 watts day, 200 watts night.

(*f*) The use of these frequencies by stations in the Amateur Service shall not cause harmful interference to the Loran system of radio navigation. If an amateur station causes such interference, the station licensee shall, as directed by the Commission, immediately cease operation on the frequencies involved.

(*g*) The use of these frequencies by the Amateur Service shall not be a bar to expansion of the radio navigation (Loran) service, and such use, and the limitations and conditions of such use as set forth above, shall be considered temporary in the sense that they shall remain subject to cancellation or to revision, in whole or in part, without hearing, whenever the Commission shall deem such cancellation or revision to be necessary or desirable in the light of the priority within this band of the Loran system of radio navigation.

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-

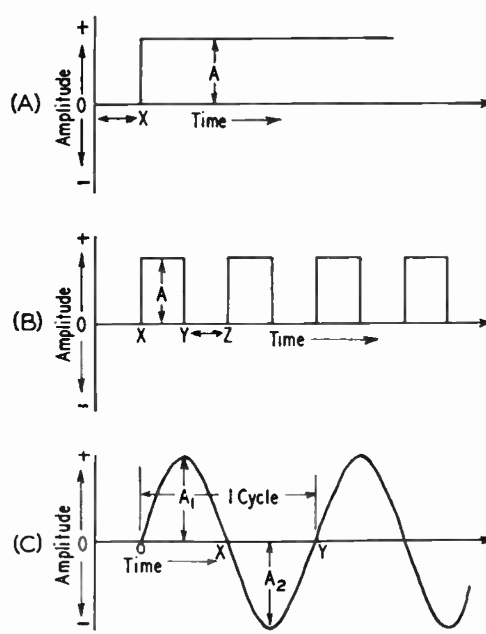


Fig. 2-1 — Three types of current flow. A — direct current; B — intermittent direct current; C — alternating current.

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

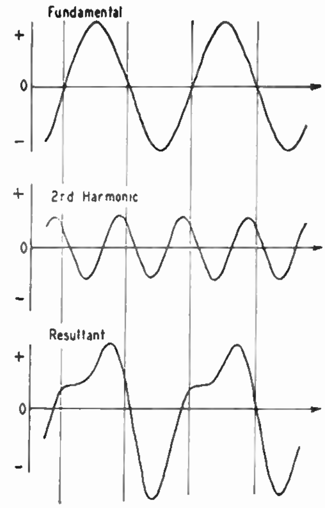


Fig. 2-2 — A complex waveform. A fundamental (top) and second harmonic (center) added together, point by point at each instant, result in the waveform shown at the bottom. When the two components have the same polarity at a selected instant, the resultant is the simple sum of the two. When they have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

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 λ = Wavelength in meters
f = Frequency in kilocycles

or

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

where λ = 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.

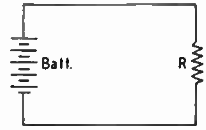


Fig. 2-3 — A simple circuit consisting of a battery and resistor.

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:

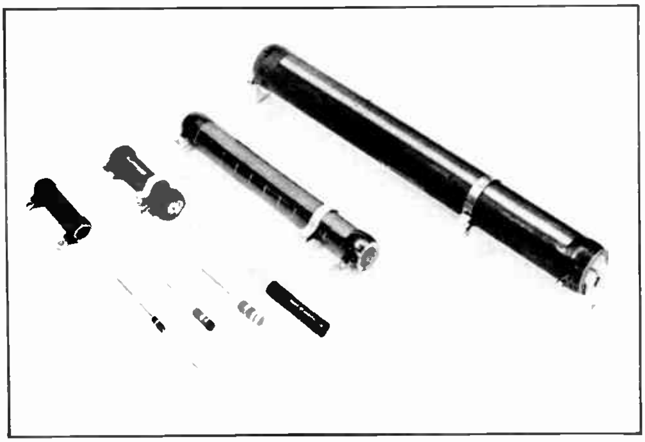
$$E = IR$$

(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 μ)
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.

TABLE 2-II
Conversion Values for Fractional and Multiple Units

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

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.

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

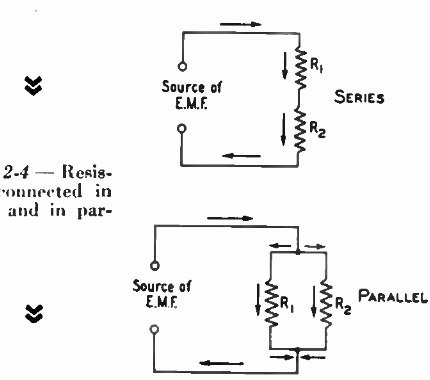


Fig. 2-4 — Resistors connected in series and in parallel.

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

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

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

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

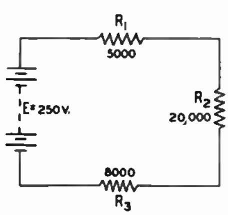


Fig. 2-5 — An example of resistors in series. The solution of the circuit is worked out in the text.

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

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

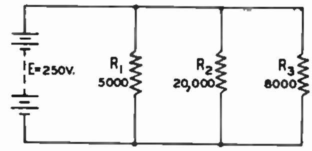


Fig. 2-6 — An example of resistors in parallel. The solution is worked out in the text.

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

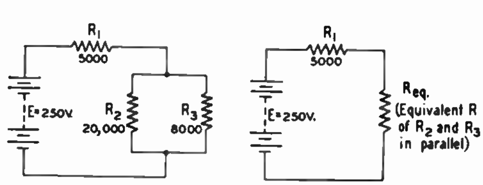


Fig. 2-7 — An example of resistors in series-parallel. The solution is worked out in the text.

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 closely enough 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.)

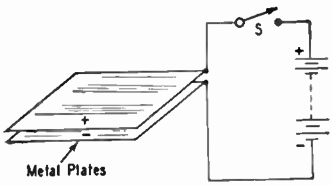


Fig. 2-8 — A simple condenser.

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).

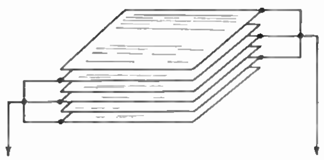


Fig. 2-9 — A multiple-plate condenser. Alternate plates are connected together.

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 **μfd.**) 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.

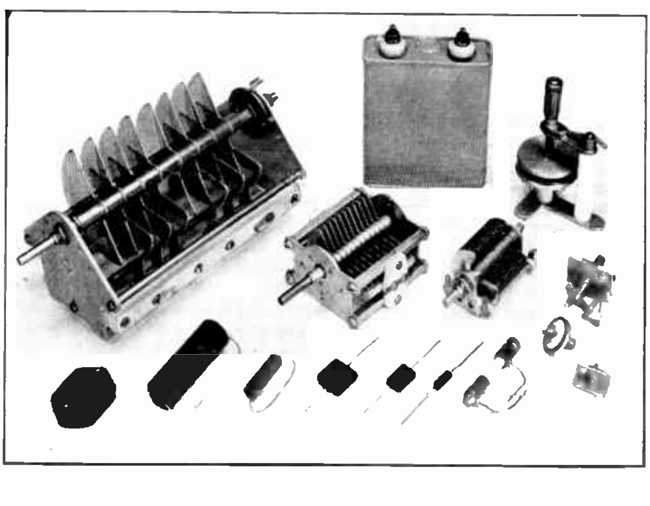
Example: A "variable" condenser has 7 semicircular plates on its rotor, the diameter of the semicircle being 2 inches. The stator has 6 rectangular plates, with a semicircular cut-out to clear the rotor shaft, but otherwise large enough to face the entire area of a rotor plate. The diameter of the cut-out is $\frac{1}{2}$ inch. The distance between the adjacent surfaces of rotor and stator plates is $\frac{1}{8}$ 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.

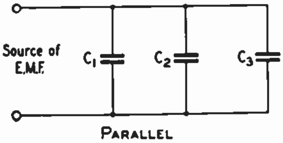
Voltage Breakdown

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



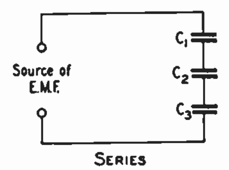


Fig. 2-10 — Condensers in series and parallel.

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

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

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,

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

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 μfd. 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.

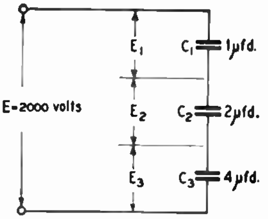


Fig. 2-11 — An example of condensers connected in series. The solution to this arrangement is worked out in the text.

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:

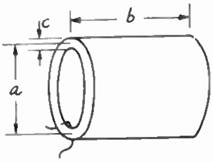


Fig. 2-12 — Coil dimensions used in the inductance formula.

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

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,

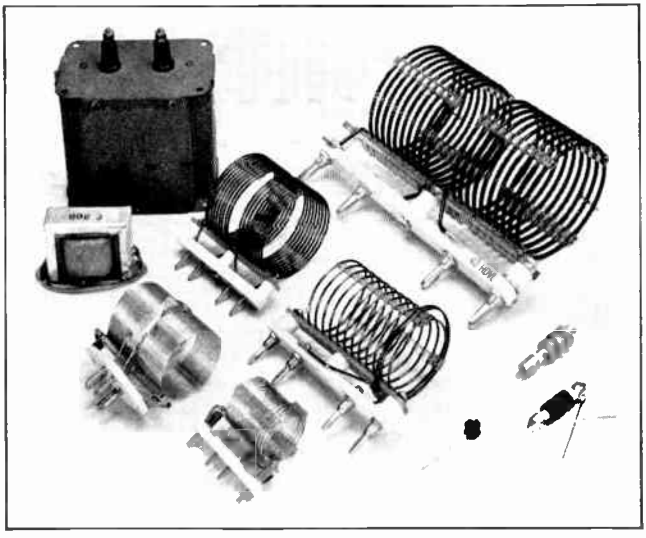
$$N = \sqrt{\frac{(3 \times 1) + (9 \times 1.25)}{0.2 \times 1^2} \times 10}$$

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

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

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-

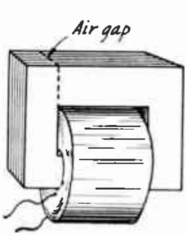


Fig. 2-13 — Typical construction of an iron-core coil. The small air gap prevents magnetic saturation of the iron and increases the inductance at high currents.

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,

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

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

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

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.

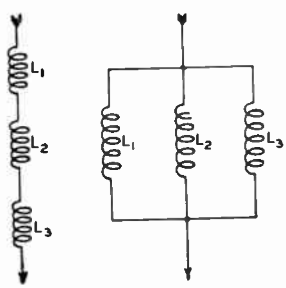


Fig. 2-14 — Inductances in series and parallel.

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

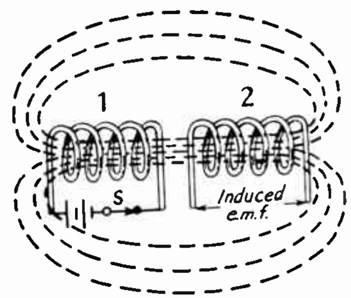


Fig. 2-15 — Mutual inductance. When the switch, *S,* is closed current flows through coil No. 1, setting up a magnetic field that induces an e.m.f. in the turns of coil No. 2.

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.

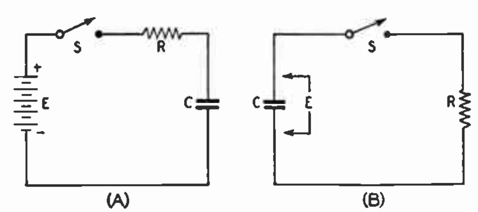


Fig. 2-16 — Schematics illustrating the time constant of an *RC* circuit.

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

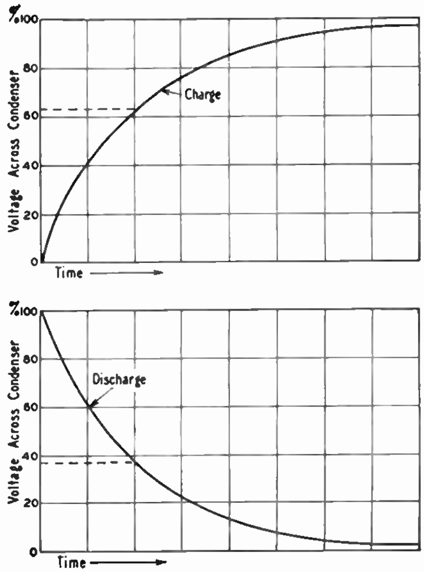


Fig. 2-17 — **How the voltage across a condenser rises, with time, when a condenser is charged through a resistor. The lower curve shows the way in which the voltage decreases across the condenser terminals on discharging through the same resistor.**

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-μfd. 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,

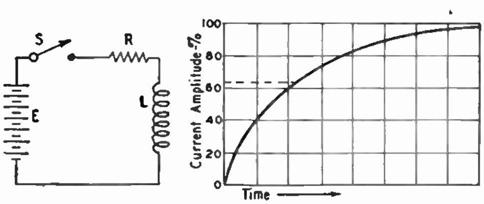


Fig. 2-18 — **Time constant of an *LR* circuit.**

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

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.

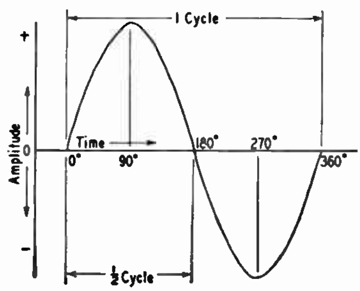


Fig. 2-19 — An a.c. cycle is divided off into 360 degrees that are used as a measure of time or phase.

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.

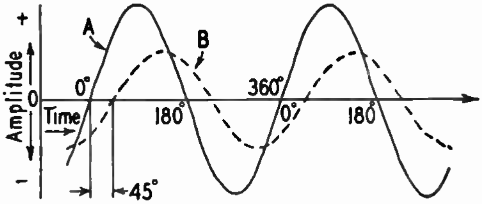


Fig. 2-20 — When two waves of the same frequency start their cycles at slightly different times, the time difference or phase difference is measured in degrees. In this drawing wave *B* starts 45 degrees (one-eighth cycle) later than wave *A*, and so lags 45 degrees behind *A*.

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.

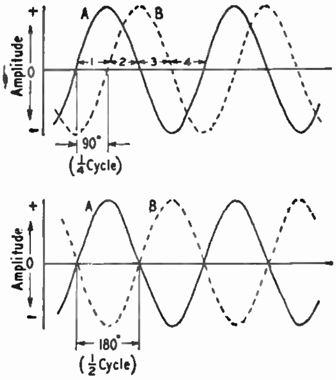


Fig. 2-21 — Two important special cases of phase difference. In the upper drawing, the phase difference between *A* and *B* is 90 degrees; in the lower drawing the phase difference is 180 degrees.

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.*

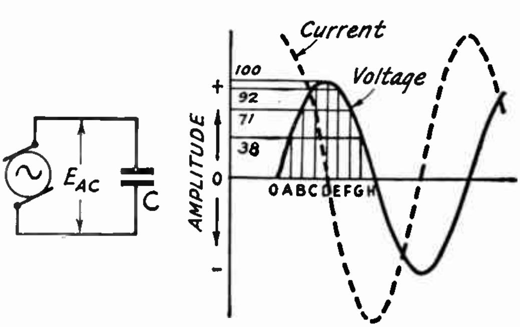


Fig. 2-22 — Voltage and current phase relationships when an alternating voltage is applied to a condenser.

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.

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

$$X_C = \frac{1}{2\pi fC}$$

where X_C = Condenser reactance in ohms
f = Frequency in cycles per second
C = Capacitance in farads
π = 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 μfd.) at a frequency of 7150 kc. (7.15 Mc.) is

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

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.

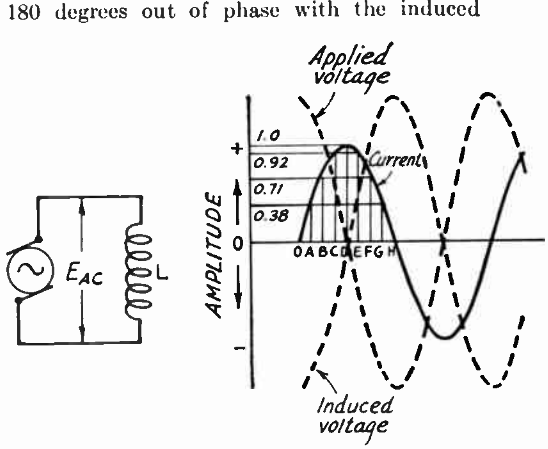


Fig. 2-23 — Phase relationships between voltage and current when an alternating voltage is applied to an inductance.

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

$$X_L = 2\pi fL$$

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

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

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

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

$X_L = 2\pi fL = 6.28 \times 14 \times 15 = 1319$ 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

$E = IX = 2 \times 47.4 = 94.8$ 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

$I = \frac{E}{X} = \frac{400}{6029} = 0.0663$ 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

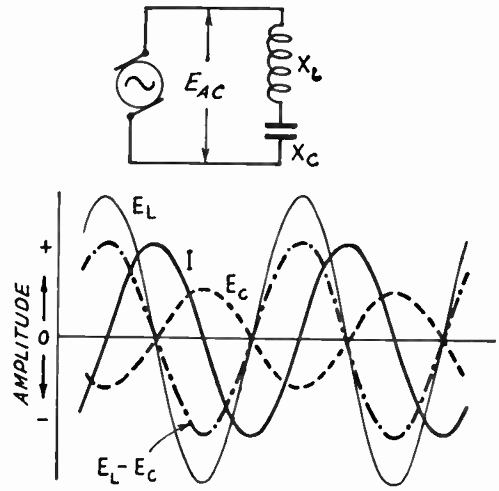


Fig. 2-24 — Current and voltages in a circuit having inductive and capacitive reactances in series.

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.

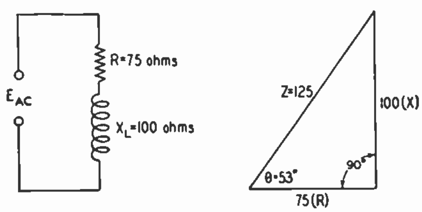


Fig. 2-25 — Resistance and inductive reactance connected in series.

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 θ 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

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

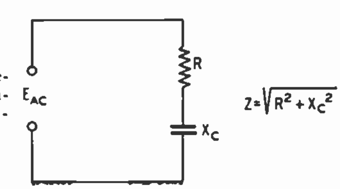
and the phase angle again is the angle whose tangent is equal to X/R. It must be remembered, however, that in this case the current *leads* the applied e.m.f., while in the resistance-inductance case it *lags* behind the voltage.

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* 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.

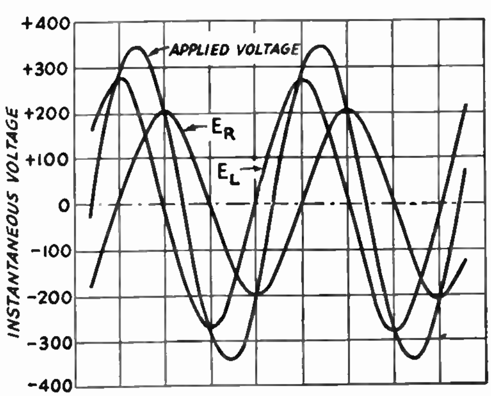


Fig. 2-27 — Voltage drops around the circuit of Fig. 2-25. Because of the phase relationships, the applied voltage is less than the arithmetical sum of the drops across the resistor and inductor.

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

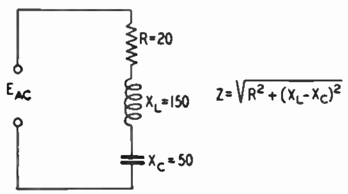


Fig. 2-28 — Resistance, inductive reactance, and capacitive reactance in series.

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

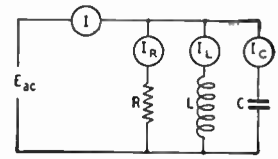


Fig. 2-29 — **Resistance, inductance and capacitance in parallel. Instruments connected as shown will read the total current, *I*, and the individual currents in the three branches of the circuit.**

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

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

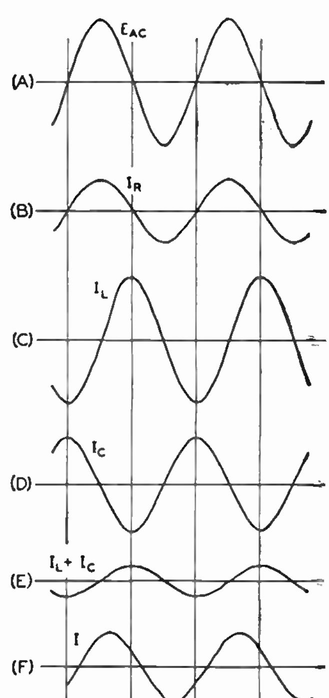


Fig. 2-30 — **Phase relationships between branch currents and applied voltage for the circuit of Fig. 2-29. The total current through *L* and *C* in parallel ($I_L + I_C$) and the total current in the entire circuit (*I*) also are shown.**

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-

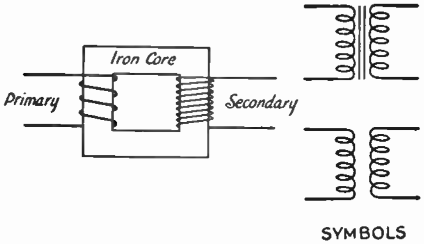


Fig. 2-31 — The transformer. Power is transferred from the primary coil to the secondary by means of the magnetic field. The upper symbol at right indicates an iron-core transformer, the lower one an air-core transformer.

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

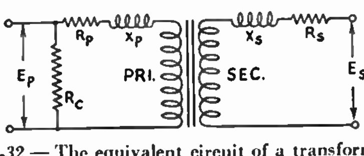


Fig. 2-32 — The equivalent circuit of a transformer includes the effects of leakage inductance and resistance of both primary and secondary windings. The resistance *Rc* is an equivalent resistance representing the constant core losses. Since these are comparatively small, their effect may be neglected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

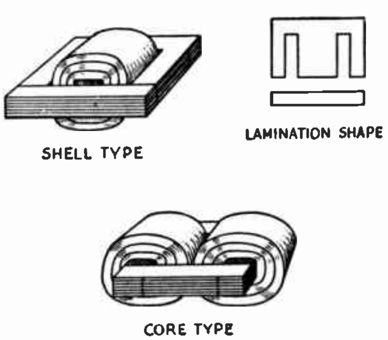


Fig. 2-33 — Two common types of transformer construction. Core pieces are interleaved to provide a continuous magnetic path with as low reluctance as possible.

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.

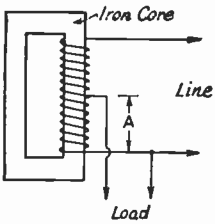


Fig. 2-34 — The autotransformer is based on the transformer principle, but uses only one winding. The line and load currents in the common winding (*A*) flow in opposite directions, so that the resultant current is the difference between them. The voltage across *A* is proportional to the turns ratio.

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 overlap 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 six to eight 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 current in the common section (A) of the winding is the difference between the line (primary) and the load (secondary) currents, since these currents are out of phase. Hence if the line and load currents are nearly equal 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

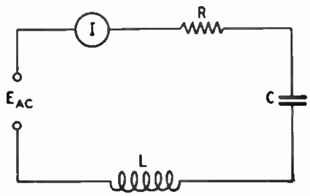


Fig. 2-35 — A series circuit containing L, C and R is "resonant" at the applied frequency when the reactance of C is equal to the reactance of L.

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
π = 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 (μh.)
C = Capacitance in micromicrofarads ($\mu\mu$fd.)
π = 3.14

Example: The resonant frequency of a series circuit containing a 5-μh. 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

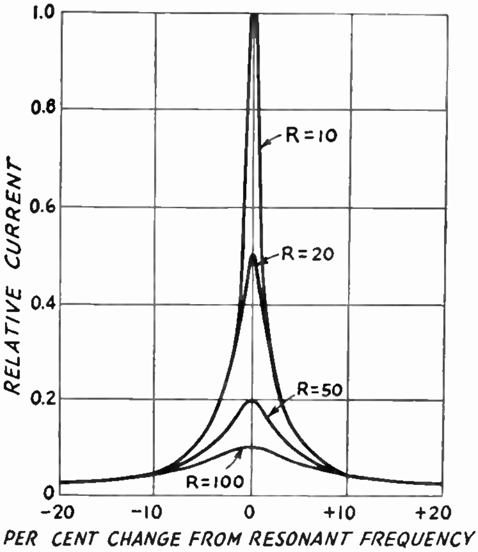


Fig. 2-36 — Current in a series-resonant circuit with various values of series resistance. The values are arbitrary and would not apply to all circuits, but represent a typical case. It is assumed that the reactances (at the resonant frequency) are 1000 ohms (minimum Q = 10). Note that at frequencies 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 higher the Q, the greater the amount of energy stored in the circuit as compared with the energy lost or used up in each cycle.

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. Qs of 10, 20, 50 and 100 are shown; these values cover much of the range commonly used in radio work.

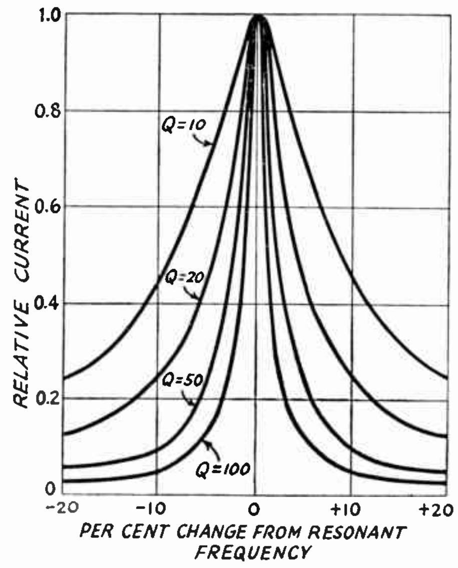


Fig. 2-37 — Current in series-resonant circuits having different Qs. In this graph the current at resonance is assumed to be the same in all cases. The lower the Q, the more slowly the current decreases as the applied frequency is moved away from resonance.

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

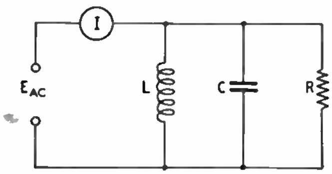


Fig. 2-38 — Circuit illustrating parallel resonance.

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.

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.)

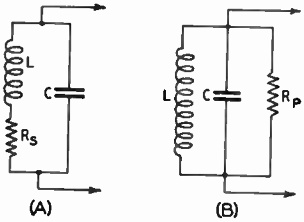


Fig. 2-39 — Series and parallel equivalents when the two circuits are resonant. The series resistor, R_s, in A can be replaced by an equivalent parallel resistor, R_p, in B, and vice versa.

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 Qs. (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-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

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

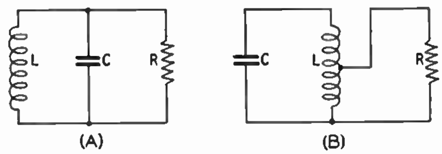


Fig. 2-41 — The equivalent circuit of a resonant circuit delivering power to a load. The resistor R represents the load resistance. At B the load is tapped across part of L, which by transformer action is equivalent to using a higher load resistance across the whole circuit.

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.

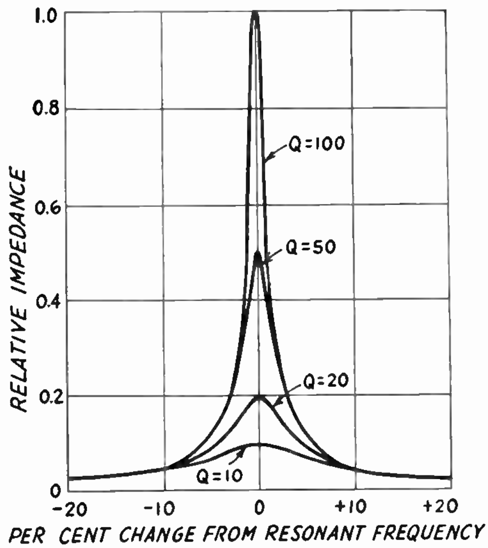


Fig. 2-40 — Relative impedance of parallel-resonant circuits with different Qs. These curves are similar to those in Fig. 2-37 for current in a series-resonant circuit. The effect of Q on impedance is most marked near the resonant frequency.

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 (μh.)
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. $L = 1900/C = 1900/25 = 76$ μh.
50 $\mu\mu$fd. $L = 1900/C = 1900/50 = 38$ μh.
100 $\mu\mu$fd. $L = 1900/C = 1900/100 = 19$ μh.
500 $\mu\mu$fd. $L = 1900/C = 1900/500 = 3.8$ μh.

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.

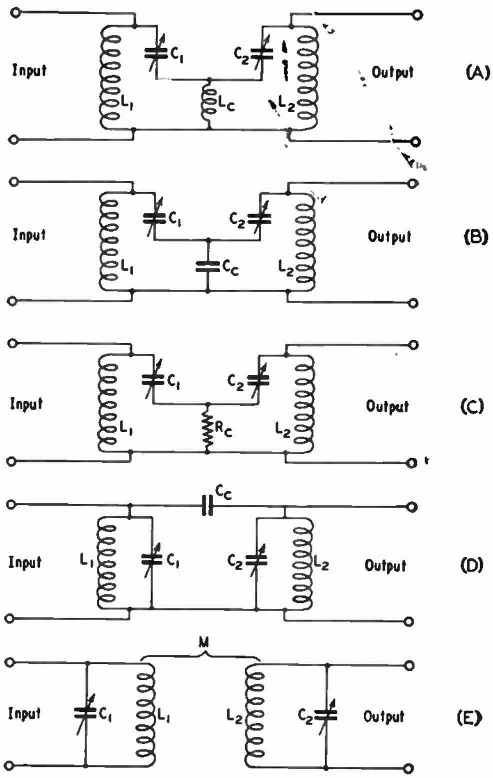


Fig. 2-42 — Basic methods of circuit coupling.

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.

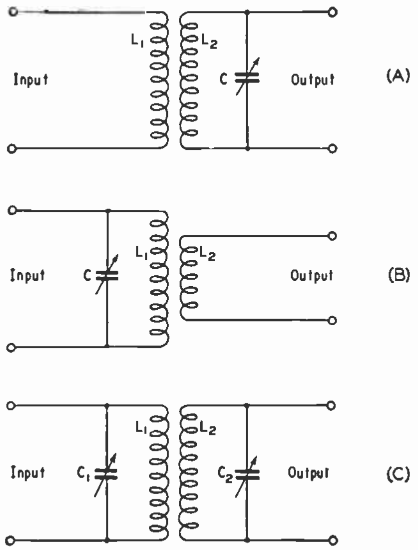


Fig. 2-43 — Types of inductively-coupled circuits. In A and B, one circuit is tuned, the other untuned. C shows the method of coupling between two tuned circuits.

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 Qs of the two circuits taken independently. A higher coefficient of coupling is required to reach critical coupling when the Qs are low; if the Qs 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

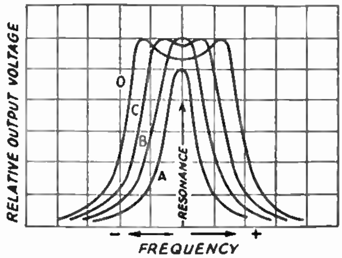


Fig. 2-44 — Showing the effect on the output voltage from the secondary circuit of changing the coefficient of coupling between two resonant circuits independently tuned to the same frequency. The voltage applied to the primary is held constant in amplitude while the frequency is varied, and the output 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 Qs 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 Qs of the individual circuits — *if* the coupling is well below critical and both circuits are tuned to resonance. The Qs 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 Qs 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.

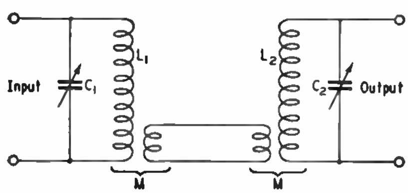


Fig. 2-45 — Link coupling. The mutual inductances at both ends of the link are equivalent to mutual inductance between the tuned circuits, and serve the same purpose.

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 Qs obtainable with LC resonant circuits.

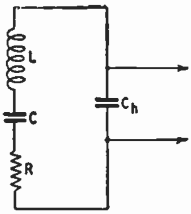


Fig. 2-46 — **Equivalent circuit of a crystal resonator. L, C and R are the electrical equivalents of mechanical properties of the crystal; C_h is the capacitance of the electrodes with the crystal plate between them.**

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

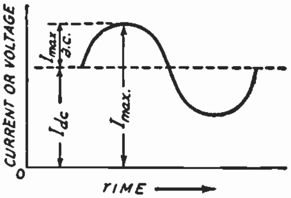


Fig. 2-47 — Pulsating current, composed of an alternating current or voltage superimposed on a steady direct current or voltage.

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.

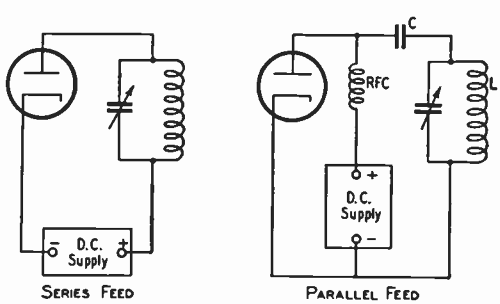


Fig. 2-48 — Illustrating series and parallel feed.

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.

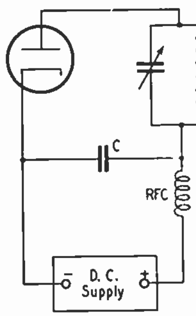


Fig. 2-49 — Typical use of a by-pass condenser in a series-feed circuit.

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 μfd. 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.

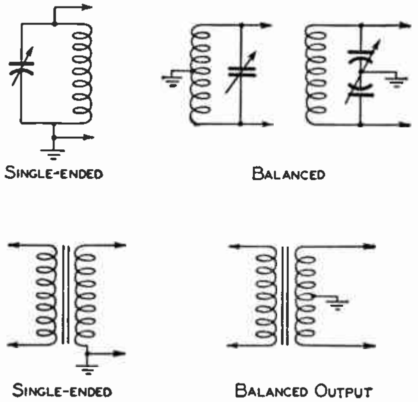


Fig. 2-50 — Single-ended and balanced circuits.

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.

CHAPTER 3

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.

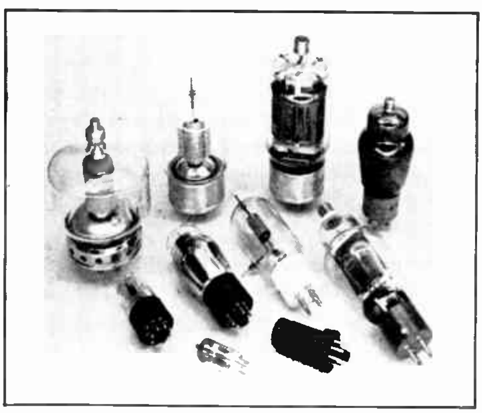
The most practical way to introduce a sufficiently-large number of electrons into the evacuated space is by **thermionic emission.**

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.

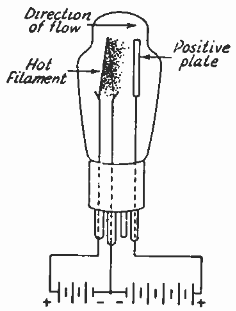


Fig. 3-1 — Conduction by thermionic emission in a vacuum tube. One battery is used to heat the filament to a temperature that will cause it to emit electrons. The other battery makes the plate positive with respect to the filament, thereby causing the emitted electrons to be attracted to the plate. Electrons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

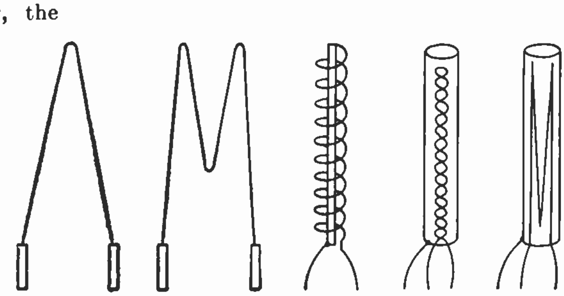


Fig. 3-2 — Types of cathode construction. Directly-heated cathodes or filaments are shown at A, B, and C. The inverted V filament is used in small receiving tubes, the M in both receiving and transmitting tubes. The spiral filament is a transmitting-tube type. The indirectly-heated cathodes at D and 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.

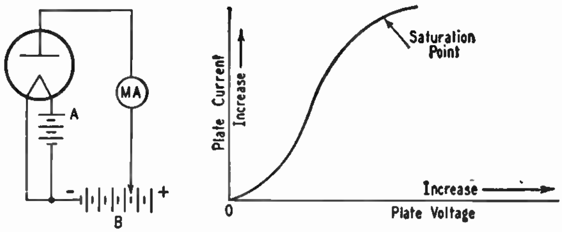
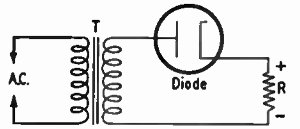


Fig. 3-3 — The diode, or two-element tube, and a typical curve showing how the plate current depends upon the voltage applied to the plate.

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.



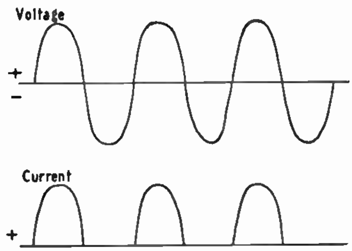


Fig. 3-4 — Rectification in a diode. Current flows only when the plate is positive with respect to the cathode, so that only half-cycles of current flow through the load resistor, *R*.

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

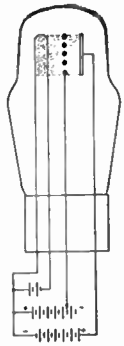


Fig. 3-5 — Construction of an elementary triode vacuum tube, showing the filament, grid (with an end view of the grid wires) and plate. The relative density of the space charge is indicated roughly by the dot density.

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

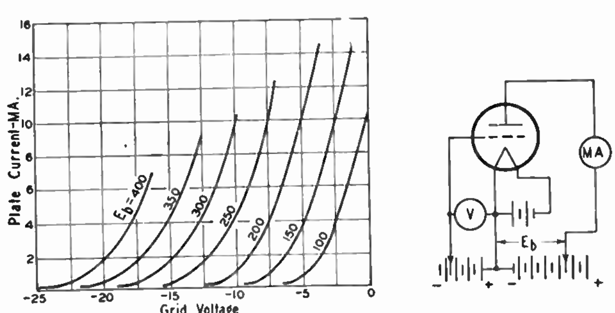


Fig 3-6 — Grid-voltage-*vs.*-plate-current curves at various fixed values of plate voltage (E_b) for a typical small triode. Characteristic curves of this type can be taken by varying the battery voltages in the circuit at the right.

to have a high **amplification factor.** Amplification factor is commonly designated by the Greek letter μ. 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-μ** tube is one with an amplification factor of perhaps 30 or more; **medium-μ** tubes have amplification factors in the approximate range 8 to 30, and low-μ tubes in the range below 7 or 8.

It would be natural to think that a tube that has a large μ would be the best amplifier, but such is not necessarily the case. If the μ 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-μ tube is relatively low. Whether or not a high-μ tube is better than one with a low μ 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

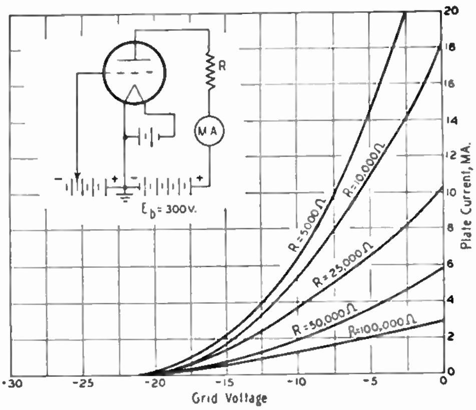


Fig. 3-7 — Dynamic characteristics of a small triode with various load resistances from 5000 to 100,000 ohms.

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 × 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

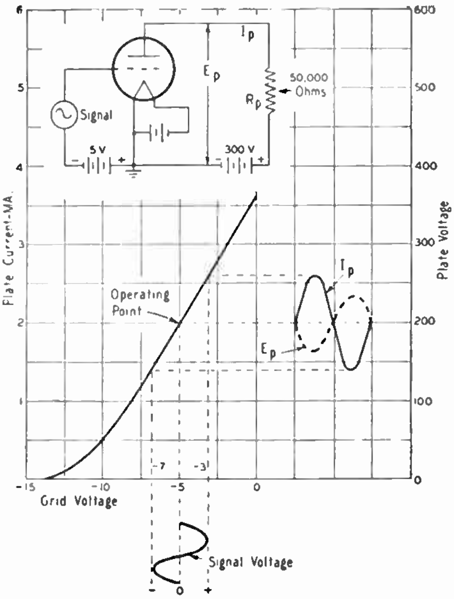


Fig. 3-8 — **Amplifier operation. When the plate current varies in response to the signal applied to the grid, a varying voltage drop appears across the load, R_p, as shown by the dashed curve, E_p. I_p is the plate current.**

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

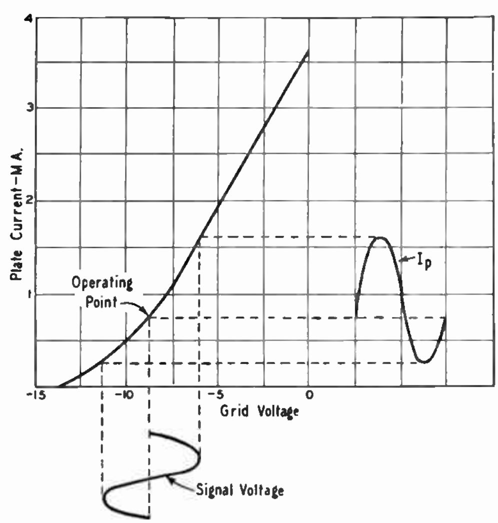


Fig. 3-9 — **Harmonic distortion resulting from choice of an operating point on the curved part of the tube characteristic. The lower half-cycle of plate current does not have the same shape as the upper half-cycle.**

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-μfd. condenser will be amply large for the usual range of audio frequencies.

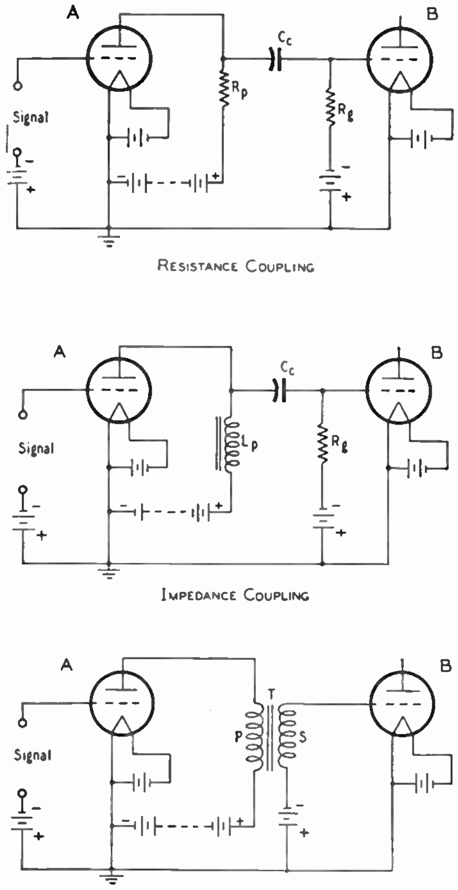


Fig. 3-10 — Three basic forms of coupling between vacuum-tube amplifiers.

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

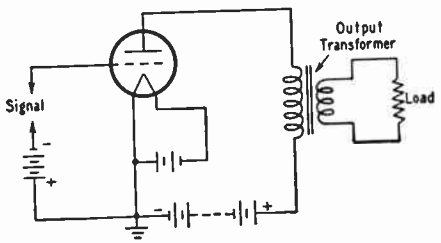


Fig. 3-11 — An elementary power-amplifier circuit in which the power-consuming load is coupled to the plate circuit through an impedance-matching transformer.

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.

An amplifier in which voltage gain is the primary consideration is called a **voltage amplifier.** Maximum voltage gain is secured when the load resistance or impedance is made as high as possible in comparison with the plate resistance of the tube. In such a case, the major portion of the voltage generated will appear across the load and only a relatively small part will be "lost" in the plate resistance.

Voltage amplifiers belong to a group called **Class A amplifiers.** A Class A amplifier is one operated so that the waveshape of the output voltage is the same as that of the signal voltage applied to the grid. If a Class A amplifier is biased so that the grid is always negative, even with the largest signal to be handled by the grid, it is called a **Class A_1 amplifier.** Voltage amplifiers are always Class A_1 amplifiers, and their primary use is in driving a following Class A_1 amplifier.

Power Amplifiers

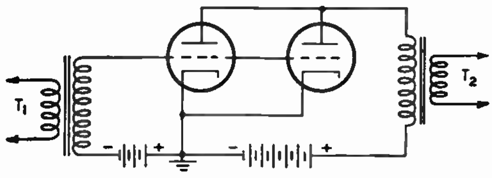
The end result of any amplification is that the amplified signal does some *work.* As a familiar example, an audio-frequency amplifier usually drives a loudspeaker that in turn produces sound waves. The greater the amount of a.f. *power* supplied to the 'speaker, the louder the sound it will produce.

In some amplifiers, therefore, *power* output rather than voltage is the primary consideration. 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. 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.

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_1 amplifier, so such an amplifier has an infinitely large power-amplification ratio. However, it is quite possible to operate a Class A amplifier in such a way that current flows in its grid circuit during at least part of the cycle. In such a case power is used up in the grid circuit and the power amplification ratio is not infinite. A tube operated in this fashion is known as a **Class A_2 amplifier.** It is necessary to use a power amplifier to drive a Class A_2 amplifier, because a voltage amplifier cannot deliver power without serious distortion of the waveshape.

Another term used in connection with power



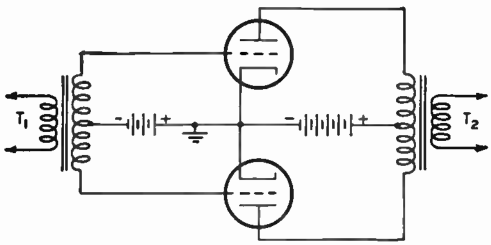


Fig. 3-12 — Parallel and push-pull a.f. amplifier circuits.

amplifiers is **power sensitivity.** In the case of a Class A_1 amplifier, it means the ratio of power output to the grid signal voltage that causes it. If grid current flows, the term usually means the ratio of plate power output to grid power input.

The a.c. power that is delivered to a load by an amplifier tube has to be paid for in power taken from the source of plate voltage and current. In fact, there is always more power going into the plate circuit of the tube than is coming out as useful output. The difference between the input and output power is used up in heating the plate of the tube, as explained previously. If we want a great deal of power output, therefore, it is advantageous to make this difference as small as possible. The ratio of useful power output to d.c. plate input is called the **plate efficiency.** The higher the plate efficiency, the greater the amount of power that can be taken from a tube having a fixed plate-dissipation rating.

Parallel and Push-Pull

When it is necessary to obtain more power output than one tube is capable of giving, two or more similar 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 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 increase successively 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 overall amplification becomes too great, there is danger that some of the output voltage will get back into one of the early stages. This "feedback," discussed in a later section, may make the amplifier unstable and prevent it from functioning as it should.

Class B Amplifiers

Fig. 3-13 shows two tubes connected in a push-pull circuit. If the grid bias is set at the point where (when no signal is applied) the plate current is just cut off, then a signal can cause plate current to flow in either tube *only* when the signal voltage applied to that particular tube is positive. In the balanced grid circuit, the signal voltages on the grids of the two tubes always have opposite polarities; that is, when the signal swings the instantaneous voltage in the positive direction on the grid of tube *A*, it is at the same time swinging the grid of tube *B* more negative. On the next half-cycle the polarities reverse and the grid of tube *B* is more positive and that of tube *A* more negative. Since the fixed bias is just at the cut-off point, this means that plate current flows only in one tube at a time.

The graphs show the operation of such an amplifier. The plate current of tube *B* is drawn inverted to show that it flows in the opposite direction, through the primary of the output transformer, to the plate current of tube *A*. Thus each half of the output-transformer primary works alternately to induce a half-cycle of voltage in the secondary. In the secondary of T_2, the original waveform is restored. This type of operation is called **Class B amplification.**

The Class B amplifier is considerably more efficient than the Class A amplifier. Furthermore, the d.c. plate current of a Class B amplifier is proportional to the signal voltage on the grids, so the power input is small with small signals. The d.c. plate power input to a Class A amplifier is the same whether the signal is large, small, or absent altogether; therefore the maximum input that can be applied to a Class A amplifier is the rated plate dissipation of the tube or tubes. Two tubes in a Class B amplifier can deliver approximately twelve times as much audio power as the same two tubes in a Class A amplifier.

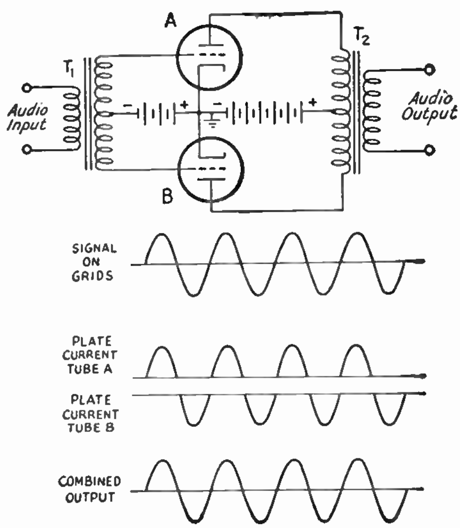


Fig. 3-13 — Class B amplifier operation.

A Class B amplifier usually is operated in such a way as to secure the maximum possible power output. This requires that the grids be driven positive with respect to the cathode during at least part of the cycle, so grid current flows and the grid circuit consumes power. While the power requirements are fairly low (as compared with the power output), the fact that the grids are positive during only *part* of the cycle means that the load on the "driver" stage varies in magnitude during the cycle; the effective load resistance is high when the grids are not drawing current and relatively low when they do take current. This must be allowed for when designing the driver.

Certain types of tubes have been designed specifically for Class B service and can be operated without fixed or other form of grid bias ("zero-bias" tubes). The amplification factor is so high that the plate current is small without signal. Because there is no fixed bias, the grids start drawing current immediately whenever a signal is applied, so the grid-current flow is continuous throughout the cycle. This makes the load on the driver much more constant than is the case with tubes of lower μ biased to plate-current cut-off.

Class AB Amplifiers

A **Class AB amplifier** is one operated midway between Class A and Class B conditions. A Class AB amplifier is a push-pull amplifier with higher bias than would be normal for pure Class A operation, but less than the cut-off bias required for Class B. At low signal levels the tubes operate practically as Class A amplifiers, and the plate current is the same with or without signal. At higher signal levels, the plate current of one tube is cut off during part of the *negative* cycle of the signal applied to its grid, and the plate current of the other tube rises with the signal. The plate current for the whole amplifier also rises above the no-signal level when a large signal is applied.

In a properly-designed Class AB amplifier the distortion is as low as with a Class A stage, but the efficiency and power output are considerably higher than with pure Class A operation. A Class AB amplifier can be operated either with or without driving the grids into the positive region. A **Class AB_1 amplifier** is one in which the grids are never positive with respect to the cathode; therefore, no driving power is required — only voltage. A **Class AB_2 amplifier** is one that has grid-current flow during part of the cycle, when the applied signal is large; it takes a small amount of driving power. The Class AB_2 amplifier will deliver somewhat more power (using the same tubes) but the Class AB_1 amplifier avoids the problem of designing a driver for it that will deliver power, without distortion, into a load of highly-variable resistance.

Class C Amplifiers

Inspection of Fig. 3-13 shows that either of the two tubes actually is working for only half the a.c. cycle and idling during the other half. It is convenient to describe the amount of time during which plate current flows in terms of electrical degrees. In Fig. 3-13 each tube has "180-degree" excitation, a half-cycle being equal to 180 degrees. The number of degrees during which plate current flows is called the **operating angle** of the amplifier. From the descriptions given above, it should be clear that a Class A amplifier has 360-degree excitation, because plate current flows during the whole cycle. In a Class AB amplifier the operating angle is between 180 and 360 degrees (in each tube) depending on the particular operating conditions chosen. The greater the amount of negative grid bias, the smaller the operating angle becomes.

An operating angle of less than 180 degrees obviously would lead to a considerable amount of distortion, because there is no way for the tube to reproduce even a half-cycle of the signal on its grid. Using two tubes in push-pull, as in Fig. 3-13, would not overcome this distortion; it would merely put together two distorted half-cycles. An operating angle of less than 180 degrees therefore cannot be used if distortionless output is wanted.

However, in certain types of amplifiers distortion does not matter particularly. One example is an amplifier used to generate r.f. power. The power output of such an amplifier is delivered to a tuned circuit, and it is characteristic of a tuned circuit that it will have a high impedance at the frequency to which it is resonant, but low impedance to all other frequencies. The tuned circuit can be made to have a high impedance at the frequency applied to the grid of the amplifier, thus providing a load of the optimum value for the tube. At harmonics of this fundamental frequency the impedance of the tuned circuit will be low, and thus will be a poor load for the tube for those frequencies set up by distortion; the distortion is "filtered out." The result is that the output voltage and current are practically pure sine waves.

Using an operating angle less than 180 degrees increases the plate efficiency, because it is characteristic of tube operation that the smaller the time during which plate current flows the smaller the amount of power lost in the plate. Also, when the proper angle and other operating conditions are chosen the power output of the amplifier is proportional to the square of the voltage applied to its plate. That is, the amplifier has the linear characteristics of a resistor insofar as its behavior when the plate voltage is varied is concerned. This is an important consideration when the amplifier is to be "modulated," as described in Chapter Nine. Such an amplifier is called a **Class C amplifier.** In Class C operation the operating angle usually is in the range 120–150 degrees, and the plate efficiency is 70 to 80 per cent.

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. 3-14 gives degenerative feed-back. Resistor R_c is in series with the regular plate resistor, R_p, and 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. 3-14 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

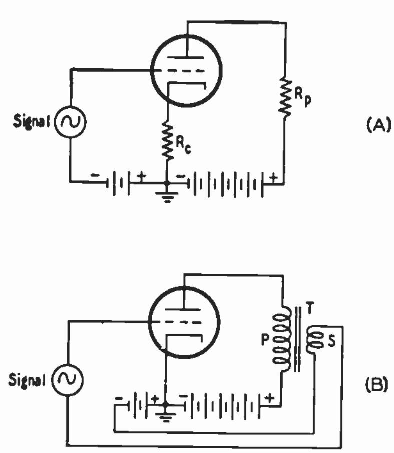


Fig. 3-14 — Circuits for producing feed-back. In A, part of the a.c. plate voltage appears across the cathode resistor, R_c, and is therefore also applied between grid and cathode. The feed-back is negative in this case. In B, the voltage that is generated in the secondary of the transformer is inserted in series in the grid circuit. Feed-back may be either positive or negative, depending upon the transformer connections.

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. 3-14 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. 3-15. 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.

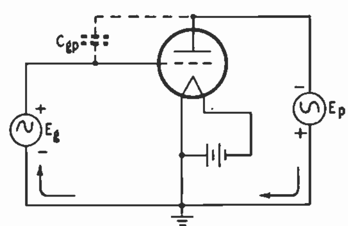


Fig. 3-15 — The a.c. voltage appearing between the grid and plate of the amplifier is the sum of the signal voltage and the output voltage, as shown by this simplified circuit. Instantaneous polarities are indicated.

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.

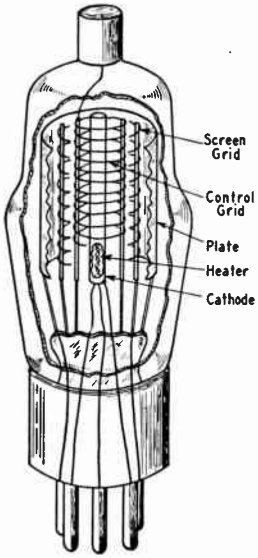


Fig. 3-16 — **Representative arrangement of elements in a screen-grid tube, with front part of plate and screen grid cut away. In this drawing the control-grid connection is made through a cap on the top of the tube, thus eliminating the capacitance that would exist between the plate- and grid-lead wires 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. 3-16. 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 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. 3-17 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

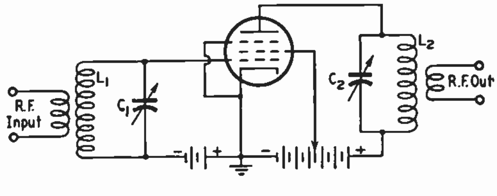


Fig. 3-17 — **Simplified pentode r.f.-amplifier circuit. L_1C_1 and L_2C_2 are tuned to the same frequency.**

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.

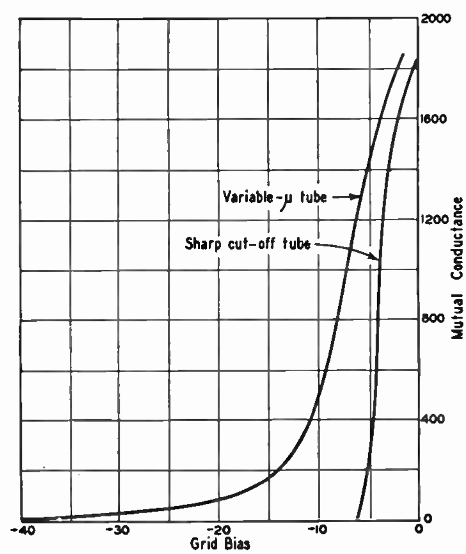


Fig. 3-18 — **Curves showing the relationship between mutual conductance and negative grid bias for two small receiving pentodes, one a sharp cut-off type and the other a variable-μ type.**

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 negative bias is increased, as shown in Fig. 3-18. 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-μ** characteristic (that is, the amplification factor changes with the grid bias), resulting in the type of curve shown in Fig. 3-18. It is evident that the variable-μ 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. 3-19. 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

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 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

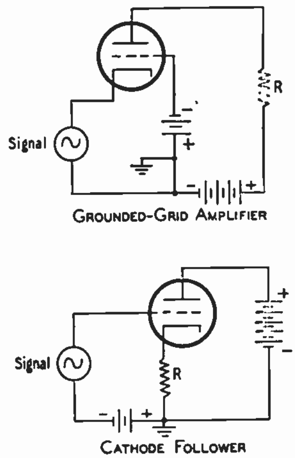


Fig. 3-19 — In the upper circuit, the grid is the junction point between the input and output circuits. In the lower drawing, the plate is the junction. In either case the output is developed in the load resistor, *R*, and may be coupled to a following amplifier by the usual methods.

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; that is, 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. 3-20. 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 employed for the most part in transmitting power tubes, where the amount of hum introduced is extremely small in comparison to the power-output level.

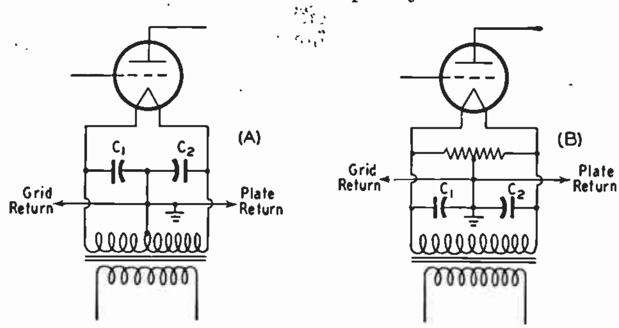


Fig. 3-20 — Filament center-tapping methods for use with directly-heated tubes.

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. 3-20.

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. 3-21. 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. 3-14A). 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 microfarads (electrolytic condensers are used for this purpose). At radio frequencies, capacitances of about 100 $\mu\mu$fd. to 0.1 μfd. 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 μfd. 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. Hence these two currents must be added when calculating the value of cathode resistor required for a screen-grid tube.

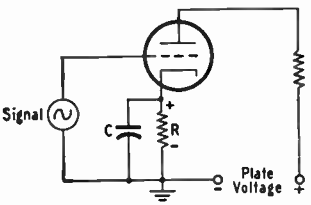


Fig. 3-21 — Cathode biasing. R is the cathode resistor and C is the cathode by-pass condenser.

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. 3-22. 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 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.

Multipurpose Tubes

A number of "combination" tubes is 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.

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.

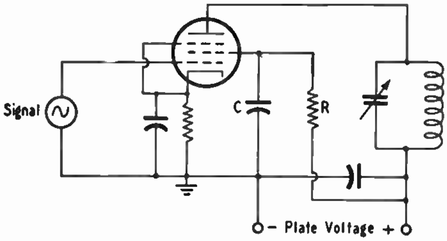


Fig. 3-22 — Screen-voltage supply for a pentode tube through a dropping resistor, R. The screen by-pass condenser, C, must have low enough reactance to bring the screen to ground potential for the frequency or frequencies being amplified.

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. 3-23A 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.

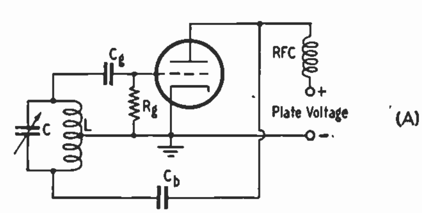
The *amount* of feed-back depends on the *position* of the tap. If the tap is too close to either end of the coil the circuit will not oscillate. If the tap is too near the grid end the voltage drop is too small to give enough feed-back, and if it is too near the plate end the impedance between the cathode 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. 3-23A 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.

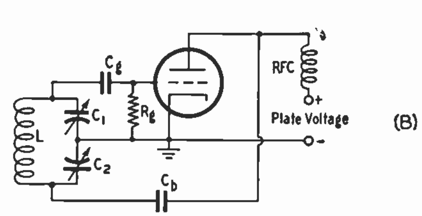
Condenser C_g is the **grid condenser.** It and R_g (the **grid leak**) are used for the purpose of obtaining grid bias for the tube. In this (and practically all) oscillator circuits the tube generates its own bias. When the grid end of the tuned circuit is positive with respect to the cathode, the grid attracts electrons from the cathode. These electrons cannot flow through L back to the cathode because C_g "blocks" 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. 3-23 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. 3-24. 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



HARTLEY CIRCUIT



COLPITTS CIRCUIT

Fig. 3-23 — 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.

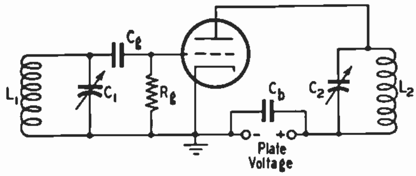


Fig. 3-24 — The tuned-plate tuned-grid oscillator.

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 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.

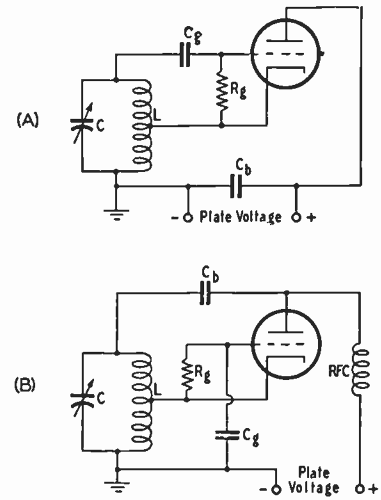


Fig. 3-25 — Showing how the r.f. ground on a typical oscillator circuit (Hartley) may be placed on either the plate (A) or grid (B) instead of the more conventional method of grounding the cathode. Provided the proper provisions are made for supplying cathode and plate voltages, the circuit operation 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. 3-23 and 3-24 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. 3-25 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. 3-23A 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. 3-25A 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.

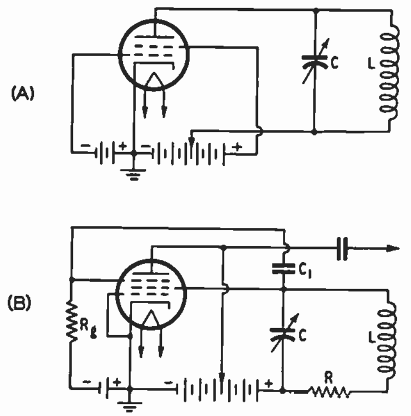


Fig. 3-26 — Negative-resistance oscillator circuits. A, dynatron; B, transitron.

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. 3-26.

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. 3-26B, 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.

CHAPTER 4

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. 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 1.8-Mc., or "160-meter," band offers reliable working over ranges up to 25 miles or so during daylight. On winter nights, ranges from 1000 to 3000 miles are not impossible. Only small sections of the band are available to amateurs, because of the presence of the loran service in that part of the spectrum. The pulse-type interference sometimes caused by loran can be readily eliminated by using an audio limiter in the receiver.

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. 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 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.

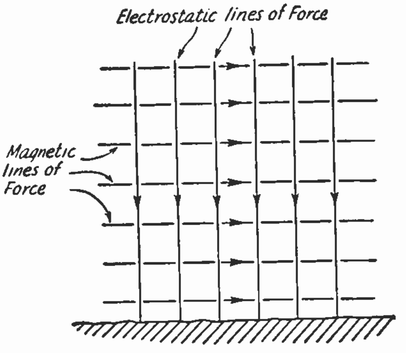


Fig. 4-1 — Representation of electrostatic and electromagnetic lines of force in a radio wave. Arrows indicate instantaneous directions of the fields for a wave traveling toward the reader. Reversing the direction of one set of lines would reverse the direction of travel.

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.

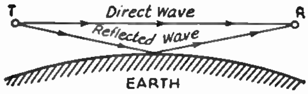


Fig. 4-2 — Showing how both direct and reflected waves may be received simultaneously in v.h.f. transmission.

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 area, the D region. The D-region 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 region 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.

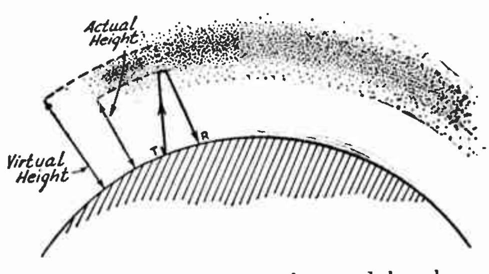


Fig. 4-3 — Bending in the ionosphere, and the echo or reflection method of determining virtual height.

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 the winter of 1947–48.

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

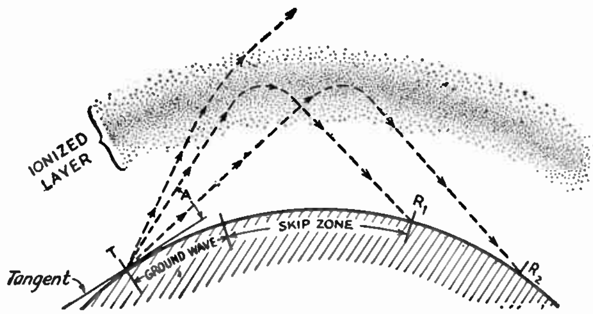


Fig. 4-4 — Refraction of sky waves, showing the critical wave angle and the skip zone. Waves leaving the transmitter at angles above the critical (greater than *A*) are not bent enough to be returned to earth. As the angle is increased, the waves return to earth at increasingly greater distances.

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.

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.

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 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 10 cents a copy or $1.00 per year on subscription. They are called "CRPL-D Basic Radio Propagation Predictions."

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 half of this. 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 2 to 4 db. have been obtained with special techniques. Comparisons of noise figures can be made by the amateur with simple equipment. (See *QST*, August, 1949, page 20.)

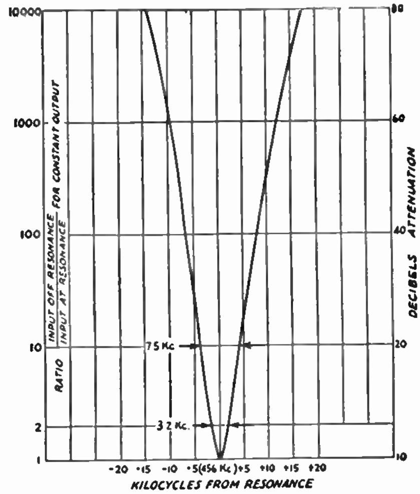


Fig. 5-1 **— Typical selectivity curve of a modern superheterodyne receiver. Relative response is plotted against deviations above and below the resonance frequency. The scale at the left is in terms of voltage ratios, the corresponding decibel steps are shown at the right.**

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

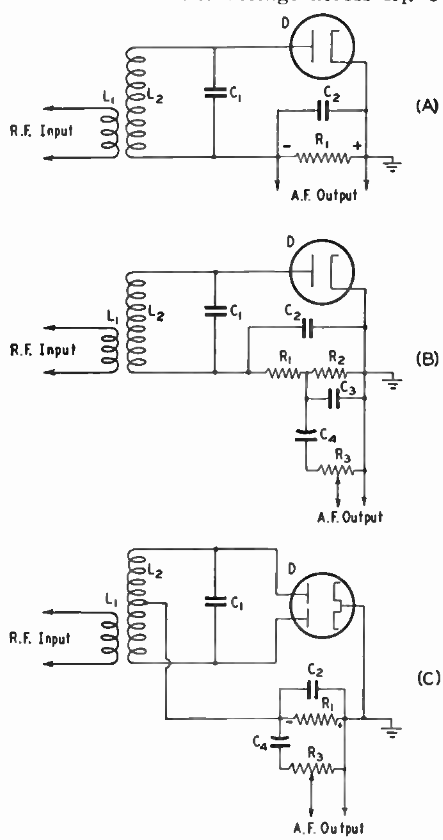


Fig. 5-2 — Simplified and practical diode detector circuits. A, the elementary half-wave diode detector; B, a practical circuit, with r.f. filtering and audio output coupling; C, full-wave diode detector, with output coupling indicated. The circuit, L_2C_1, is tuned to the signal frequency; typical values for C_2 and R_1 in A and C are 250 μμfd. and 250,000 ohms, respectively; in B, C_2 and C_3 are 100 μμfd. each; R_1, 50,000 ohms; and R_2, 250,000 ohms. C_4 is 0.1 μfd. 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.

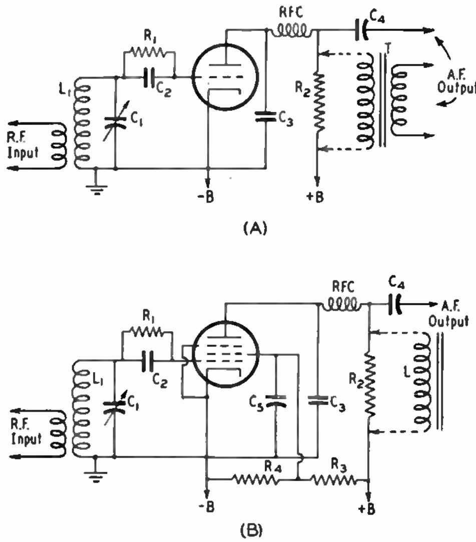


Fig. 5-4 — Grid-leak detector circuits, A, triode; B, pentode. A tetrode may be used in the circuit of B by neglecting the suppressor-grid connection. Transformer coupling may be substituted for resistance coupling in A, or a high-inductance choke may replace the plate resistor in B. L_1C_1 is a circuit 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 μμfd.	100 to 250 μμfd.
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	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		300- to 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.

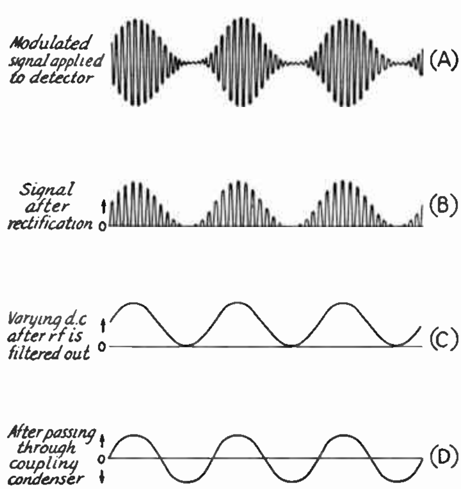


Fig. 5-3 — Diagrams showing the detection process.

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.

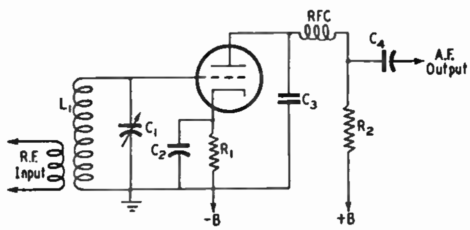
The sensitivity of the grid-leak detector is higher than that of any other type. Like the diode, it "loads" the tuned circuit and reduces its selectivity. The linearity is rather poor, and the signal-handling capability is limited. The signal-handling capability can be improved by reducing R_1 to 0.1 megohm, but the sensitivity will be decreased. The chief use of the grid-leak detector is as a regenerative detector in simple receivers.

Plate Detectors

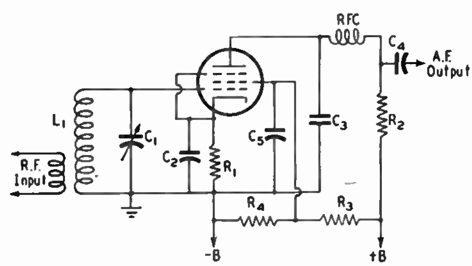
The plate detector is arranged so that rectification of the r.f. signal takes place in the plate circuit of the tube, as contrasted to the grid rectification just described. Sufficient negative bias is applied to the grid to bring the plate current nearly to the cut-off point, so that the application 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 satisfac. tory, 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 $\mu\mu$fd.
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 usually 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.

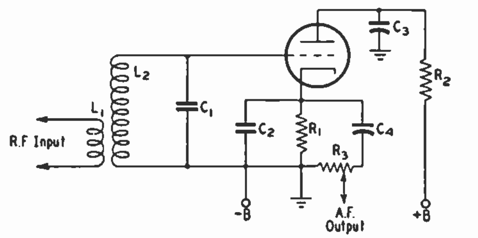


Fig. 5-6 — The infinite-impedance detector. The input circuit, L_2C_1, is tuned to the signal frequency. Typical values for the other components are:

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, generally 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 only when the circuit of Fig. 5-7C is used,

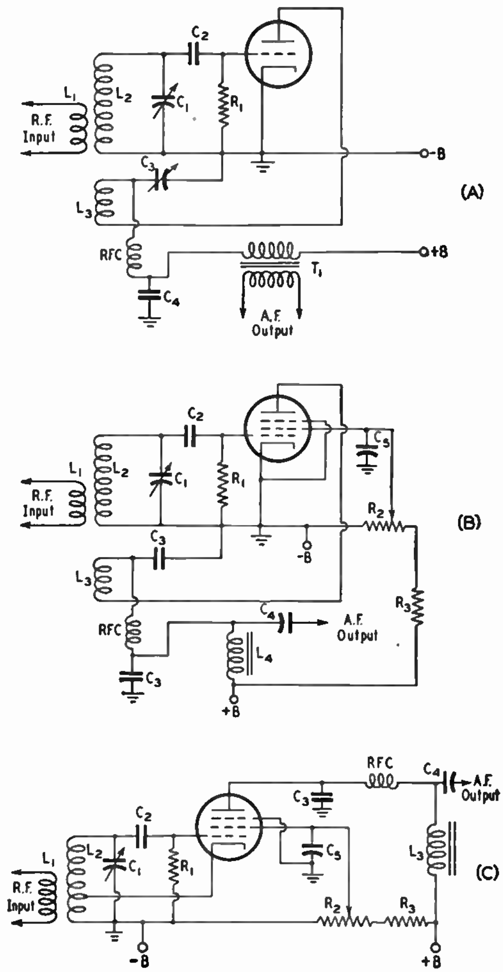


Fig. 5-7 — Triode and pentode regenerative detector circuits. The input circuit, L_2C_1, is tuned to the signal frequency. The grid condenser, C_2, should have a value of about 100 μμfd. in all circuits; the grid leak, R_1, may range in value from 1 to 5 megohms. The tickler coil, L_3, ordinarily will 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 μμfd. or more; by-pass condensers C_3 in B and C are likewise 100 μμfd. C_5 is ordinarily 1 μfd. 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 μfd. 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.

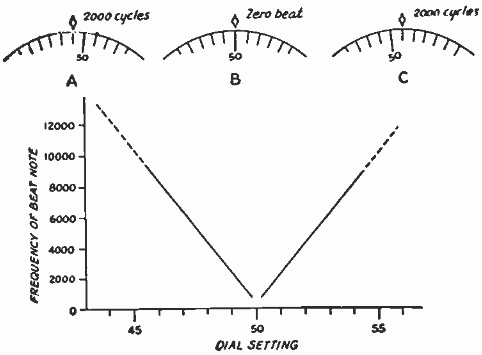


Fig. 5-8 — As the tuning dial of a receiver is turned past a c.w. signal, the beat-note varies from a high tone down through "zero beat" (no audible frequency difference) and back up to a high tone, as shown at A, B and C. The curve is a graphical representation of the action. The beat exists past 8000 or 10,000 cycles but usually is not heard because of the limitations of the audio system.

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 μfd.) 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 r.f. 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 is heard 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 sensitive to 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.

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

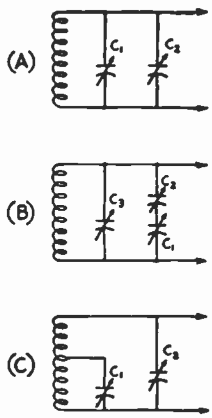


Fig. 5-9 — Essentials of the three basic bandspread tuning systems.

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. 5-9.

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. 5-10, where C_1 is the trimmer and C_2 the tuning condenser. The use of the trimmer necessarily

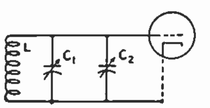


Fig. 5-10 — Showing the use of a trimmer condenser to set the minimum circuit capacity in order to obtain true tracking for gang-tuning.

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. 5-9. 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. 5-9B, and C_2 in Fig. 5-9C, 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

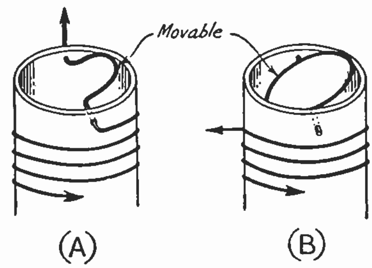


Fig. 5-11 — Methods of adjusting the inductance for ganging. The half-turn in A can be moved so that its magnetic field either aids or opposes the field of the coil. The shorted loop in B is not connected to the coil, but operates by induction. It will have no effect on the coil inductance when the axis 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. 5-11.

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. 5-12. The variations are chiefly in the way in which the oscillator voltage is introduced. In 5-12A, 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] 160[2]	18,000	22,000	0.5
6SB7Y / 6BA7	250	100	0[1] / 75[2]	12,000 / 15,000	22,000	0.35

[1]Self-excited. [2]Separate excitation.

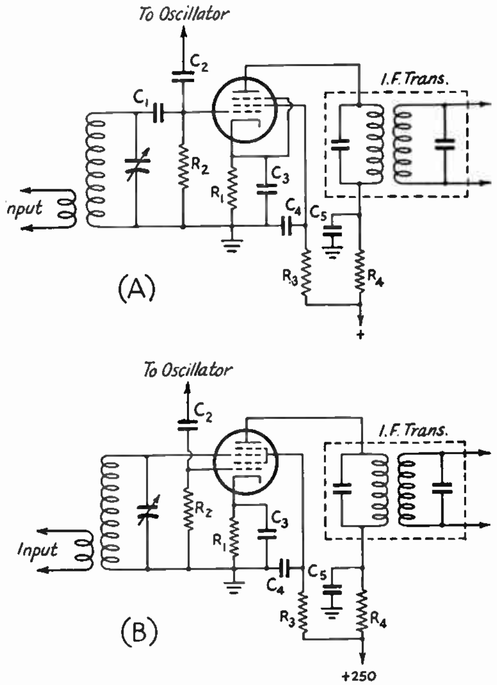


Fig. 5-12 — Typical circuits for separately-excited mixers. Grid injection of a pentode mixer is shown at A, and separate excitation of a pentagrid converter is given in B. Typical values for B will be found in Table 5-I — the values below are for the pentode mixer of A.

C_1 — 10 to 50 $\mu\mu$fd.
C_2 — 5 to 10 $\mu\mu$fd.
C_3, C_4, C_5 — 0.001 μfd.
R_1 — 6800 ohms.
R_2 — 1.0 megohm.
R_3 — 0.47 megohm.
R_4 — 1500 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), the 12AT7 (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. 5-12B, 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. 5-13. The circuit shown in Fig. 5-13A, 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. 5-13B 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 very 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. 5-12 and 5-13.

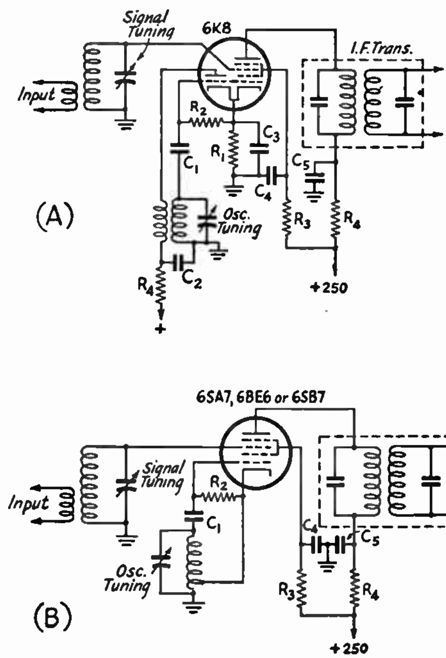


Fig. 5-13 — Typical circuits for triode-hexode (A) and pentagrid (B) converters. Values for R_1, R_2 and R_3 can be found in Table 5-I; others are given below.

C_1 — 47 $\mu\mu$fd.
C_2, C_4, C_5 — 0.001 μfd.
C_3 — 0.01 μfd.
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 mechanical shock and changes in voltage and loading. Thermal effects (slow change in frequency because of tube or circuit heating) should be minimized. They can be reduced by using ceramic instead of bakelite insulation in the r.f. circuits, a large cabinet relative to 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 can be a bother, and should 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.

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. 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.

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.

The oscillator plate voltage should be as low as is consistent with adequate output. Low plate voltage will reduce tube heating and thereby lower the 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.

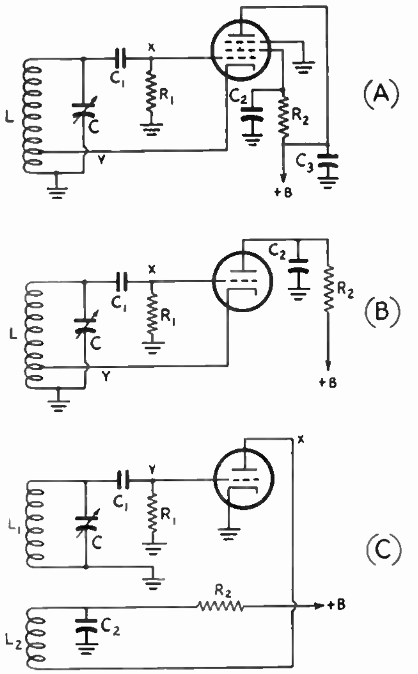


Fig. 5-14 — **High-frequency oscillator circuits. A, pentode grounded-plate oscillator; B, triode grounded-plate oscillator; C, triode oscillator with tickler circuit. Coupling to the mixer may be taken from points X and Y. In A and B, coupling from Y will reduce pulling effects, but gives less voltage 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.

Circuits

Several oscillator circuits are shown in Fig. 5-14. 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. 5-14C 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. Using the oscillator shown at C, feed-back is obtained by increasing the number of turns on L_2 or by moving L_2 closer to L_1.

The Intermediate-Frequency Amplifier

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 cut-

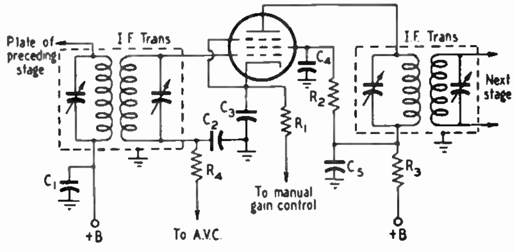


Fig. 5-15 — Typical intermediate-frequency amplifier circuit for a superheterodyne receiver. Representative values for components are as follows:

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.

ting reduces fidelity, it is frequently preferable to sacrifice naturalness of reproduction in favor of communications effectiveness.

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.

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. 5-15. 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. 5-15, 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.

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

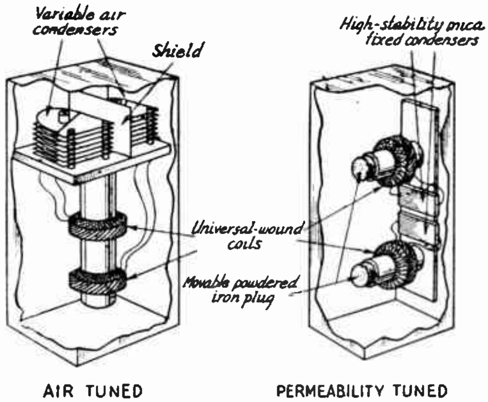


Fig. 5-16 — **Representative i.f.-transformer construction. Coils are supported on insulating tubing or (in the air-tuned type) on wax-impregnated wooden dowels. The shield in the air-tuned transformer prevents capacity coupling between the tuning condensers. In the permeability-tuned transformer the 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.**

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 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 Qs 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, 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

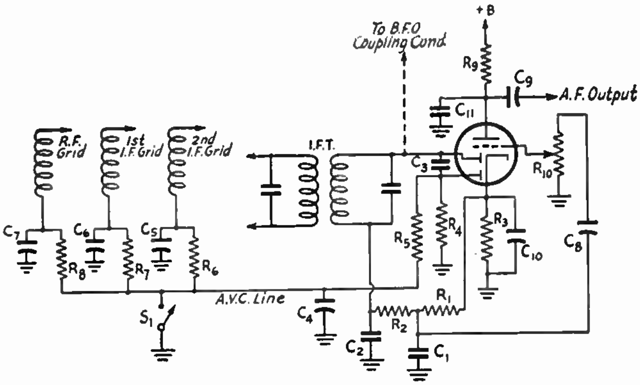


Fig. 5-17 — Automatic volume-control circuit using a dual-diode-triode as a combined a.v.c. rectifier, second detector and first a.f. amplifier.
R_1 — 0.27 megohm.
R_2 — 50,000 to 250,000 ohms.
R_3 — 1800 ohms.
R_4 — 2 to 5 megohms.
R_5 — 0.5 to 1 megohm.
R_6, R_7, R_8, R_9 — 0.25 megohm.
R_{10} — 0.5-megohm variable.
C_1, C_2, C_3 — 100 $\mu\mu$fd.
C_4 — 0.1 μfd.
C_5, C_6, C_7 — 0.01 μfd.
C_8, C_9 — 0.01 to 0.1 μfd.
C_{10} — 5- to 10-μfd. electrolytic.
C_{11} — 270 $\mu\mu$fd.

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. 5-16.

Besides the type of i.f. transformer shown in Fig. 5-16, 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.

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:

Intermediate Frequency	*Bandwidth in Kilocycles* *2 times down*	*10 times down*	*100 times down*
One stage, 50 kc. (iron core)	0.8	1.4	2.8
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-μ (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 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. 5-14A 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. 5-17. 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 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 amplifiers 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. 5-17. 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.

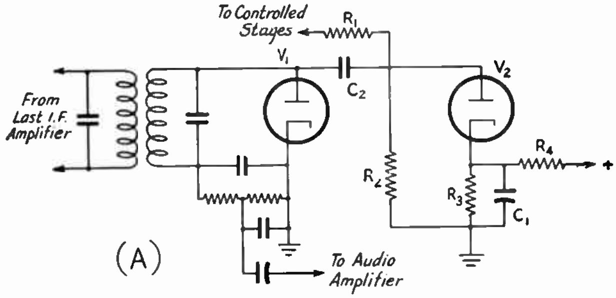
In Fig. 5-17 the audio-diode return is made directly to the cathode and the a.v.c. diode is returned to ground. This places negative bias on the a.v.c. diode equal to the d.c. drop through the cathode resistor (a volt or two) and thus delays the application of a.v.c. voltage to the amplifier grids, since 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. 5-17 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. A.v.c. systems designed to work on c.w. signals must have fairly long time constants to work with slow-speed sending, and often a selection of time constants is made available.



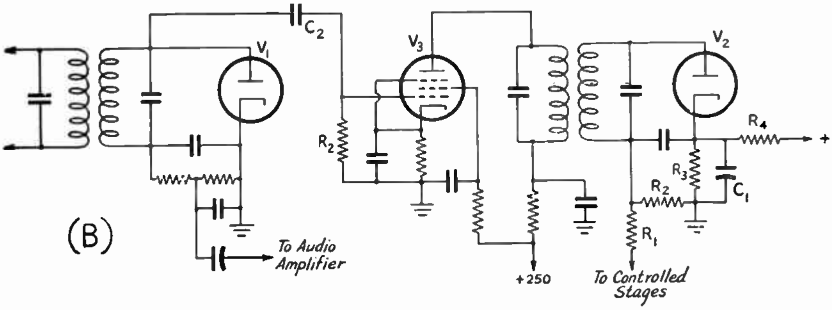


Fig. 5-18 — Delayed a.v.c. is shown at A, and amplified and delayed a.v.c. is shown in B. The circuit at B gives excellent a.v.c. action over a wide range, with no impairment of sensitivity for weak signals. For either circuit, typical values are:
C_1 — 0.001 μfd. R_1, R_2 — 1.0 megohm.
C_2 — 100 μμfd. 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 in the case of a multitube high-gain affair.

Amplified A.V.C.

The a.v.c. system shown in Fig. 5-17 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. 5-17, 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. 5-18A. 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. 5-18B. 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 with 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 with 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.

SECOND-DETECTOR NOISE-LIMITER CIRCUITS

The circuit of Fig. 5-19 "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.

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 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. 5-19A, 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,

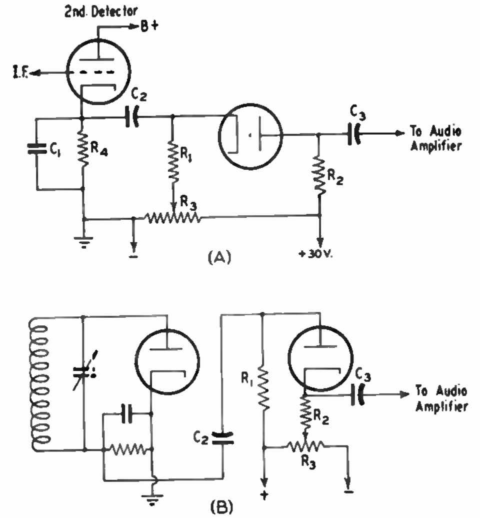


Fig. 5-19 — Series-valve noise-limiter circuits. A, as used with an infinite-impedance detector; B, with a diode detector. Typical values for components are as follows:

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 $\mu\mu$fd.
C_2, C_3 — 0.1 μfd.

All other diode-circuit constants in B are conventional.

the circuit arrangement shown in Fig. 5-19B must be used.

An audio signal of about ten volts is required for good limiting action. The limiter will work on either c.w. or 'phone signals, but in either case the potentiometer must be set at a point determined by the strength of the incoming signal.

Second-detector noise-limiting circuits that automatically adjust themselves to the receiver carrier level are shown in Fig. 5-20. 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

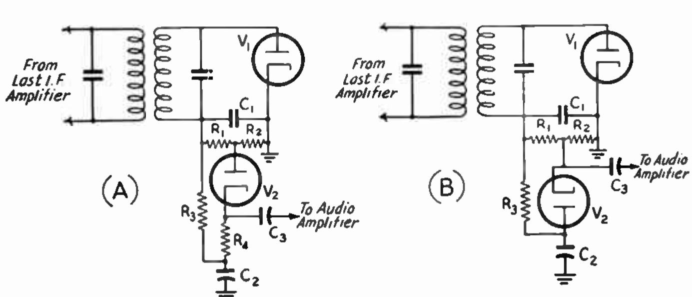


Fig. 5-20 — Self-adjusting series (A) and shunt (B) noise limiters. The functions of V_1 and V_2 can be combined in one tube like the 6H6 or 6AL5, or Type 1N34 crystals can be used.

C_1 — 100 $\mu\mu$fd.
C_2, C_3 — 0.05 μfd.
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.

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 the circuit at B will be found in the eight-tube superheterodyne 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. 5-21, 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.

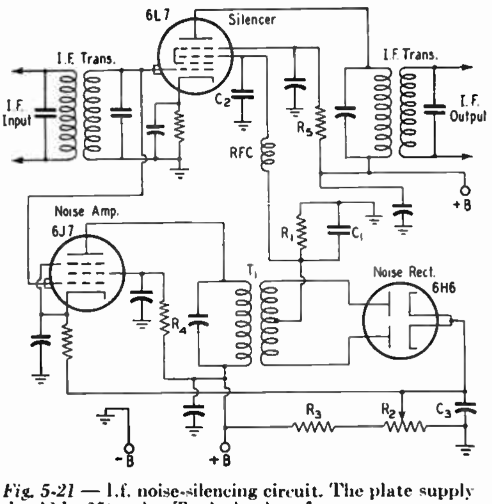


Fig. 5-21 — I.f. noise-silencing circuit. The plate supply should be 250 volts. Typical values for components are:
C_1 — 50-250 μμfd. (use smallest value possible without r.f. feed-back).
C_2 — 17 μμfd. R_2 — 5000-ohm variable.
C_3 — 0.1 μfd. R_3 — 22,000 ohms.
R_1, R_4, R_5 — 0.1 meg. RFC — 20 mh.
T_1 — Special i.f. transformer for noise rectifier.

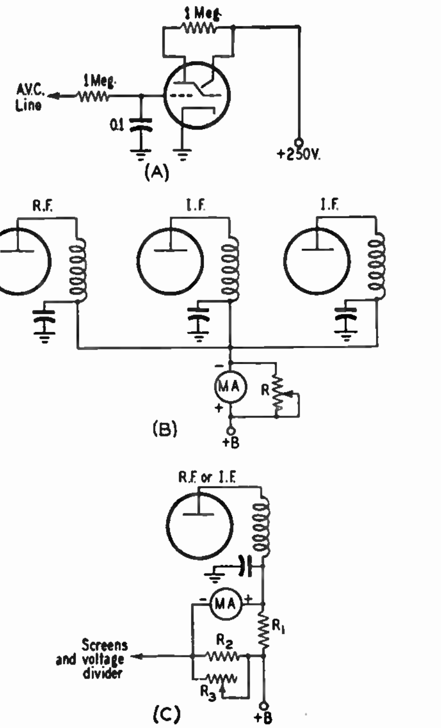


Fig. 5-22 — Tuning-indicator or S-meter circuits for superhet receivers. A, electron-ray indicator; B, plate-current meter for tubes on a.v.c.; C, bridge circuit for a.v.c.-controlled tube. In B, resistor R should have a maximum resistance several times that of the milliammeter. In C, representative values for the components are: R_1, 270 ohms; R_2, 330 ohms; R_3, 1000-ohm variable.

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.

Three types of indicators are shown in Fig. 5-22. 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. 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.

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 attenuated to a very low level.

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.

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 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 selectivity varies.

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. 5-23 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 and, by reducing the band-width, reduces the response of the receiver to noise.

Crystal-Filter Circuits; Phasing

Several crystal-filter circuits are shown in Fig. 5-24. 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 undesired signals.

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. 5-23.

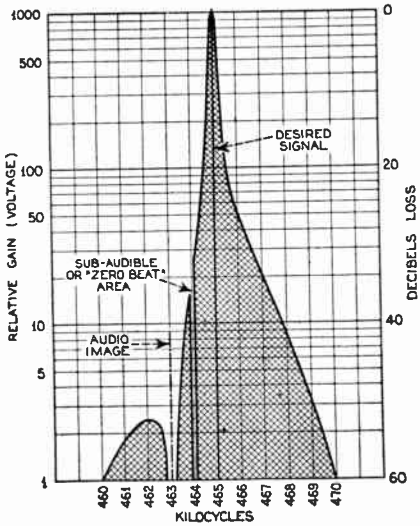


Fig. 5-23— Graphical representation of single-signal selectivity. The shaded area indicates the over-all bandwidth, or region in which response is obtainable.

Additional I.F. Selectivity

Most commercial communications receivers do not have sufficient selectivity for amateur use, and their performance can be greatly improved by adding additional selectivity. One popular method is to couple a BC-453 aircraft receiver (war surplus, tuning range 190 to 550 kc.) to the tail end of the 465-kc. i.f. amplifier in the communications receiver and use the resultant output of the BC-453. The aircraft receiver uses an 85-kc. i.f. amplifier that is quite sharp — 6.5 kc. wide at −60 db. — and it helps tremendously in separating 'phone signals and in backing up crystal filters for improved c.w. reception. (See *QST*, January, 1948, page 40.)

If a BC-453 is not available, it is still a simple matter to enjoy the benefits of improved selectivity. It is only necessary to heterodyne to a lower frequency the 465-kc. signal existing in the receiver i.f. amplifier and then rectify it after passing it through the sharp low-frequency amplifier. The Hammarlund Company and the J. W. Miller Company both offer 50-kc. transformers for this application.

QST references on high i.f. selectivity include: McLaughlin, "Selectable Single Sideband," April, 1948; Githens, "C.W. Receiver," Aug., 1948.

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.

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. (Regen-

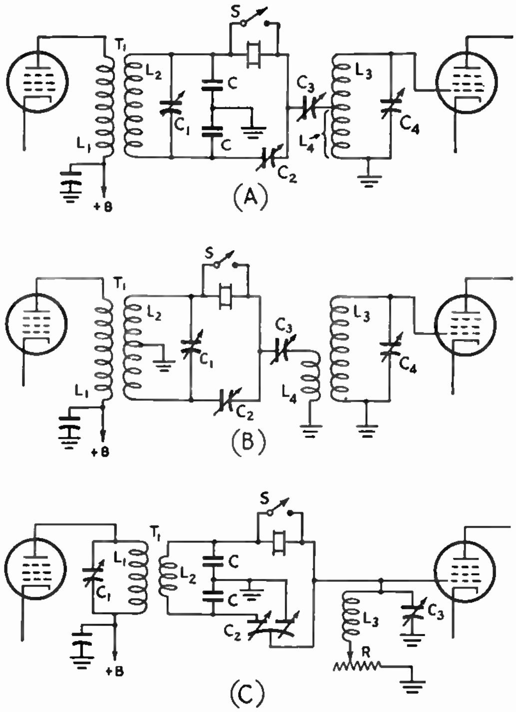


Fig. 5-24 — Crystal-filter circuits of three types. All give variable bandwidth, with C having the greatest range of selectivity. Suitable circuit values are as follows: Circuit A, T_1, special i.f. input transformer with high-inductance primary, L_1, closely coupled to tuned secondary, L_2; C_1, 50-$\mu\mu$fd. 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).

eration 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 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-μ 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.

Feed-Back

Feed-back giving rise to regeneration and oscillation 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 an 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, it is sometimes necessary to shield grid and plate leads and to be careful not to run them close together.

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, instead of resistors, are used in the heater leads where necessary.

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 an 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-μ 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-μ 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. 5-25.

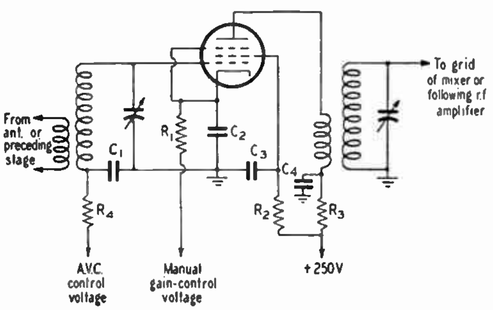


Fig. 5-25 — Typical radio-frequency amplifier circuit for a superheterodyne receiver. Representative values for components are as follows:

C_1, C_2, C_3, C_4 — 0.01 μfd. below 15 Mc., 0.001 μfd. at 30 Mc.
R_1, R_2 — See Table 5-II.
R_3 — 1800 ohms.
R_4 — 0.22 megohm.

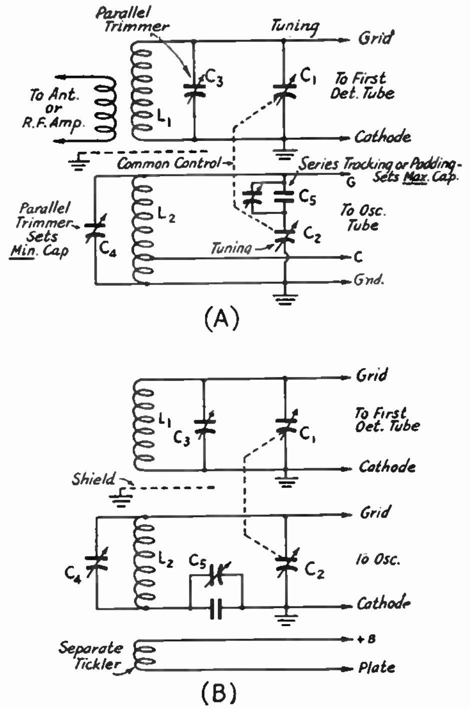


Fig. 5-26 — Converter-circuit tracking methods. Following are approximate circuit values for 450- to 465-kc. i.f.s, with tuning ranges of approximately 2.15-to-1 and C_2 having 140-$\mu\mu$fd. maximum, and the total minimum capacitance, including C_3 or C_4, being 30 to 35 $\mu\mu$fd.

Tuning Range	L_1	L_2	C_5
1.7–4 Mc.	50 μh.	40 μh.	0.0013 μfd.
3.7–7.5 Mc.	14 μh.	12.2 μh.	0.0022 μfd.
7–15 Mc.	3.5 μh.	3 μh.	0.0045 μfd.
14–30 Mc.	0.8 μh.	0.78 μh.	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_2	C_5
0.5–1.5 Mc.	240 μh.	130 μh.	425 $\mu\mu$fd.
1.5–4 Mc.	32 μh.	25 μh.	0.00115 μfd.
4–10 Mc.	4.5 μh.	4 μh.	0.0028 μfd.
10–25 Mc.	0.8 μh.	0.75 μh.	None used

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. 5-26. 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. 5-26. 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. 5-9C, 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 5-9A if the value of C_1 is small compared with C_2.

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 tne 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 around 30 Mc. The 6J4, 6J6, 7F8 and triode-connected 6AK5 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 futile to try to combine best sensitivity with 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.

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.

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 that necessary with legitimate signals. In poorly-shielded receivers it is often possible to find b.f.o. harmonics below 2 Mc., but they should be very weak at higher frequencies.

Narrow-Band Frequency- and Phase-Modulation Reception

FM Reception

In the reception of NFM 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 FCC regulations, except in those portions of the bands where wide-band FM is permitted.

If the receiver has a discriminator or other detector designed expressly for FM reception, the signal is *peaked* on the receiver (as indicated by maximum S-meter reading or minimum background noise). There is also a spot on either side of this tuning condition where audio is recovered through slope detection, but the signal will not be as loud and the background noise will be higher.

PM Reception

Phase-modulated 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 can readily be overcome by using additional audio gain. NFM signals received through the crystal filter in this fashion will have a "boomy" characteristic because the lower frequencies are accentuated.

Reception of Single-Sideband Signals

Single-sideband signals are generally transmitted with little or no carrier, and it is necessary to furnish the carrier at the receiver before proper reception can be obtained. Because little or no carrier is transmitted, the a.v.c. in the receiver is not useful, and manual variation of the r.f. gain control is required.

A single-sideband signal can be identified by the absence of a strong carrier and by the severe variation of the S-meter at a syllabic rate. When such a signal is encountered, it should first be peaked with the main tuning dial. (This centers the signal in the i.f. passband.) After this operation, do not touch the main tuning dial. Then set the r.f. gain control at a very low level and switch off the a.v.c. Increase the audio volume control to maximum, and bring up the r.f. gain control until the signal can be heard weakly. Switch on the beat oscillator, and carefully adjust the frequency of the beat oscillator until proper speech is heard. If there is a slight amount of carrier present, it is only necessary to *zero-beat* the oscillator with this weak carrier. It will be noticed that with an incorrect setting of the beat oscillator, the speech will sound high- or low-pitched or even inverted (very garbled), but no trouble will be had in getting the correct setting, once a little experience has been obtained. The use of minimum r.f. gain and maximum audio gain will insure that no distortion (overload) occurs in the receiver.

Another method of receiving single-sideband signals is to reinsert the carrier at the signal frequency. If, for example, you wish to copy a single-sideband signal that is on 3990 kc., you can supply the carrier at that frequency (with a small auxiliary oscillator or frequency meter) and leave your receiver in the normal condition for AM reception (a.v.c. on, b.f.o. off). This method of reception is advantageous in "round-table" contacts that include a single-sideband station, because it calls only for careful tuning of the auxiliary oscillator and not of the receiver. Further, only the auxiliary oscillator must be stable.

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-μfd. 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 μfd. 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. The better receivers use VR-type tubes to stabilize the oscillator voltage — a defective tube will cause trouble with oscillator instability.

A One-Tube Regenerative Receiver

The receiver shown in Figs. 5-27, 5-28, 5-29 and 5-30 represents close to the minimum requirements of a useful short-wave receiver. Under suitable conditions, it is capable of receiving signals from many foreign countries. It is an excellent receiver for the beginner, because it is easy to build and the components are not expensive.

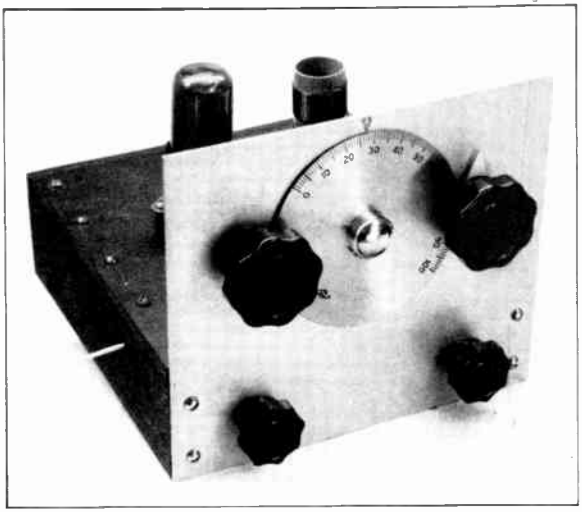


Fig. 5-27 — **The simple one-tube regenerative receiver is built on a wood-and-Presdwood chassis, with an aluminum panel. The large left-hand knob drives the calibrated scale on the bandspread condenser. The large right-hand knob is for the band-set condenser.**

From the circuit in Fig. 5-29, it can be seen that the only tube in the receiver is a 6SN7 twin triode. One section is used as a regenerative detector, the other triode section serving as an audio amplifier to the headphones. A variable antenna-coupling condenser, C_1, minimizes "dead spots" in the tuning range that might be caused by antenna-resonance effects. Two tuning condensers are used. The band-set condenser, C_4, tunes to the desired frequency band, and the bandspread condenser, C_2/C_3, allows the operator to tune slowly through the band. The bandspread condenser is a dual condenser made from a single midget variable, and on all of the amateur bands except 3.5 Mc. only the C_3 portion is connected in the circuit. The 3.5-Mc. coil includes a jumper that connects C_2 on that band. Regeneration is controlled by varying the plate voltage on the detector with R_4.

The mechanical design is made as simple as possible. Work on the chassis and the front panel can be done with only a No. 8 drill, a ¼-inch drill, and a round file. There is no complicated metal work or bending. To reduce the panel size, the knob on the band-set condenser overlaps the friction-driven tuning dial.

The front panel is a 7 × 7-inch sheet of $\frac{1}{16}$-inch aluminum. It carries the tuning controls, the regeneration adjustment and the antenna-coupling condenser shaft. The sides of the chassis are soft wood strips, 7 × 2 × ⅝ inches. The deck of the chassis is a 7 × 7-inch sheet of ¼-inch Presdwood (or Masonite). The 6SN7 socket is supported on ⅝-inch-long mounting pillars, and the 5-

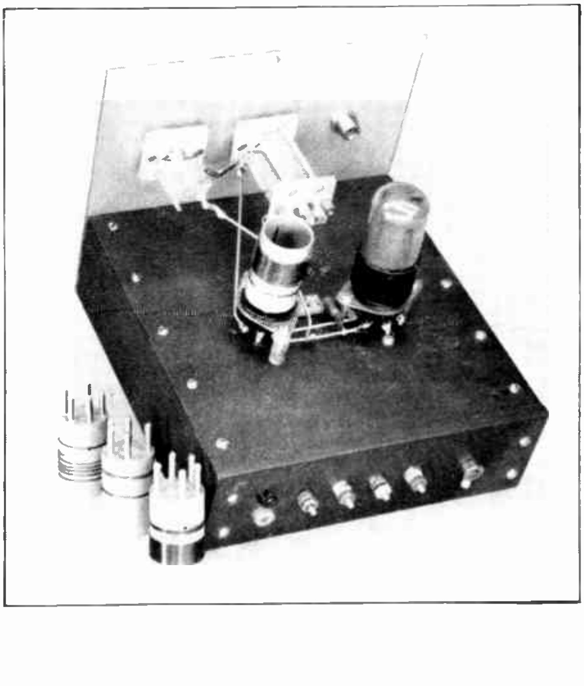


Fig. 5-28 — **Another view of the one-tube regenerative receiver shows how the tube and coil sockets are mounted. The headphone tips plug into the two small tip jacks on the rear panel — the set of four machine screws and nuts is for connecting to the power supply.**

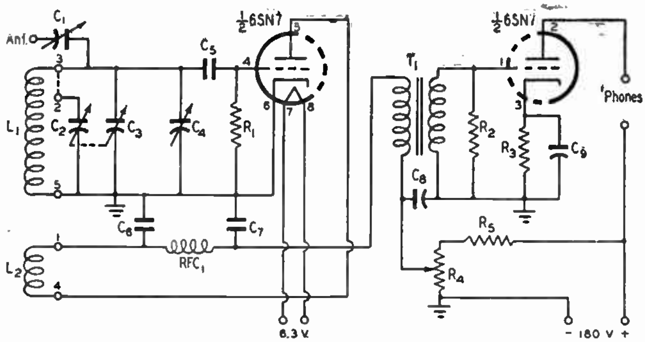


Fig. 5-29 — Wiring diagram of the one-tube regenerative receiver.

- C_1 — Homemade adjustable condenser. See text.
- C_2, C_3 — Reworked midget variable (Millen 21935). See text.
- C_4 — 100-$\mu\mu$fd. midget variable (Millen 20100).
- C_5 — 100-$\mu\mu$fd. mica.
- C_6, C_7 — 470-$\mu\mu$fd. mica.
- C_8 — 12-μfd. 150-volt electrolytic.
- C_9 — 10-μfd. 25-volt electrolytic.
- R_1 — 1.5 megohms, ½ watt.
- R_2 — 0.15 megohm, ½ watt.
- R_3 — 1500 ohms, ½ watt.
- R_4 — 50,000-ohm wire-wound potentiometer.
- R_5 — 33,000 ohms, 1 watt.
- RFC_1 — 2.5-mh. r.f. choke (National 100U).
- T_1 — Interstage audio transformer (Stancor A-4723).

prong coil socket is on ⅞-inch pillars. The grid leak, R_1, and grid condenser, C_5, are located above the deck. The back panel is made of ¼-inch Presdwood and carries the binding posts. The binding posts are ¾-inch 6–32 machine screws with suitable nuts and washers. The chassis is assembled with ¾-inch No. 6 round-head wood screws. Upon completion, the assembly is given a coat of flat black paint. The front panel is secured to the chassis side members with No. 6 round-head wood screws.

The bandspread condenser, C_2/C_3, is made by modifying a Millen 21935 variable condenser. Using a hack-saw blade, the stator bars are carefully cut between the eighth and ninth plates (counting back from the front panel). The ninth plate is removed by twisting it loose with long-nosed pliers.

Coil sizes and data are given in the coil table. All coils are wound on 1-inch diameter 5-pin coil forms. The coil for the 80-meter range is close-wound and requires no treatment, but the spaced-turns coils should be secured by running a thin line of Duco cement across the wire at several points. Before cementing the turns in place, each coil should be tried in the receiver. To obtain smooth regeneration, it may be necessary to make minor coupling adjustments (changes in spacing) between L_1 and L_2.

The antenna condenser, C_1, is made from two 1-inch squares of sheet copper. One plate is secured to the underside of the deck on a tie-point. The other plate is carried by a ¼-inch diameter polystyrene rod. Rotating the shaft swings the moving plate away from the fixed plate and provides a capacity of from 5 to less than 1 $\mu\mu$fd. The polystyrene rod passes through the front panel and out the back panel. It is secured at the back by a ¼-inch shaft collar. The panel end carries a tuning knob, and a rubber grommet under slight compression, placed between the knob and the panel, acts as a friction lock. The moving plate is secured to the polystyrene rod by a copper-wire hairpin soldered to the plate and fixed into a pair of holes drilled in the rod. A flexible

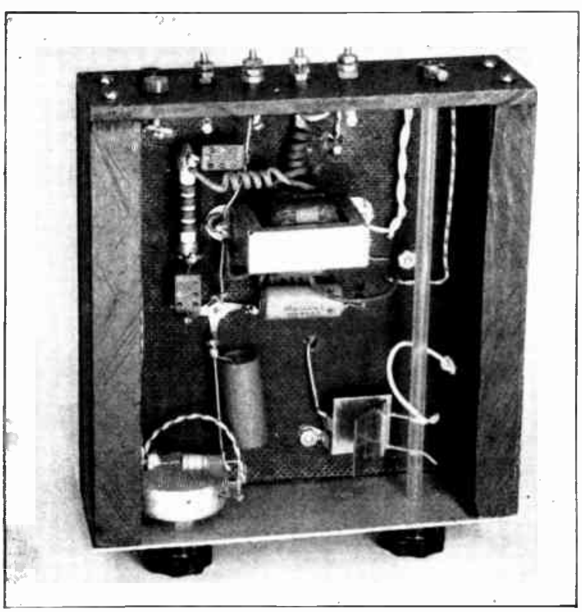


Fig. 5-30 — This view underneath the one-tube regenerative receiver shows the arrangement of parts and the construction of the variable antenna-coupling condenser.

COIL TABLE FOR THE ONE-TUBE REGENERATIVE RECEIVER

All coils wound on Millen 45005 1-inch diameter coil forms. Both L_1 and L_2 should be wound in the same direction, with L_2 closer to the pins of the form. The grid end of L_1 and the plate end of L_2 should be on the outside ends of the coils.

Range	L_1	L_2	*Sep.* L_1-L_2
2.8 — 6 Mc. (80 meters)	25 t. No. 26 enam., close-wound	4 t. No. 26 enam., close-wound	⅜ inch
5.9 — 13.5 Mc. (40 meters)	13½ t. No. 22 enam., spaced to occupy ⅝ inch	1¼ t. No. 26 enam., close-wound	¼ inch
13.6 — 30 Mc. (20 and 14 meters)	5¼ t. No. 22 enam., spaced to occupy ⅝ inch	1¾ t. No. 26 enam., close-wound	⅜ inch
24.5 — 40 Mc. (10 and 11 meters)	1½ t. No. 22 enam., close-wound	1¾ t. No. 26 enam., close-wound	5/16 inch

separation between strips is just enough (1¼ inches) to clear the tube socket and electrolytic condensers, and the leads from the transformer and choke also pass through this opening. Binding posts are made in the same manner as on the receiver, with No. 6 machine screws and suitable nuts and washers.

Although it is satisfactory to mount the power supply on the same table with the receiver, it should be at least one or two feet away, to avoid the possibility of a.c. hum pick-up. For the same reason, the antenna lead should not pass too close to any a.c. wiring from or to the power supply.

Using the parts listed in Fig. 5-32 should result in a power supply that gives about 180 volts when connected to the receiver. However, if the 6SN7 in the receiver appears to run too hot (as tested by touching the tube after the receiver has been running for 5 or 10 minutes), the output voltage can be reduced by increasing the resistance at R_1 (Fig. 5-32). Adding

lead is soldered to the protruding wire, and the lead passes out through a hole in the side of the chassis to make connection to the antenna. Knots in this wire, on either side of the chassis wall, secure the wire firmly in place. The fixed plate is covered with a single layer of cellophane Scotch Tape, to prevent a short-circuit when the condenser is positioned at maximum capacity.

All wiring is No. 14 tinned copper. Direct leads from the condensers to the coil socket add to the strength and rigidity of the receiver. The r.f. choke RFC_1, by-pass condensers, and the audio transformer all are fastened to the underside of the deck.

The power supply for the receiver, shown in Figs. 5-31 and 5-32, is simple to assemble because it is built on a wooden chassis. Two strips of 1½ × ¾-inch wood, 12 inches long, are nailed to two short end pieces. The

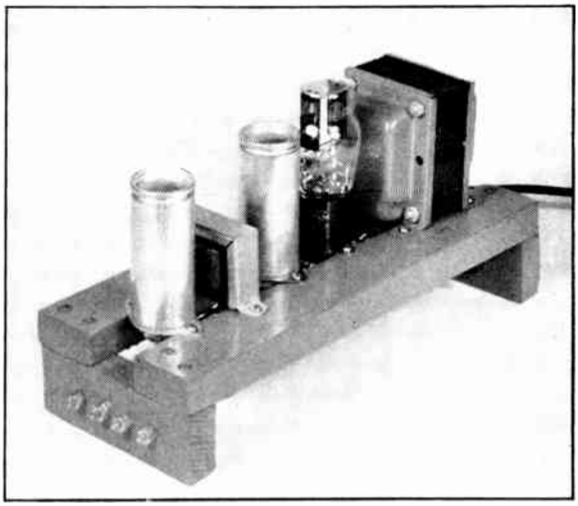


Fig. 5-31 — The power supply for the regenerative receiver is built on a simple wooden chassis.

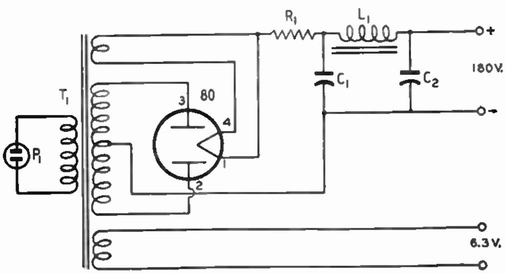


Fig. 5-32 — Circuit diagram of the power supply for the regenerative receiver.

C_1, C_2 — 16-μfd. 450-volt electrolytic (Mallory RS-217).
R_1 — 20,000-ohm 10-watt wire-wound.
L_1 — 15-henry 50-ma. filter choke (Stancor C-1080).
P_1 — 115-volt line plug.
T_1 — 275-0-275 volts at 50 ma., 6.3 v. at 2.5 amp., 5 v. at 2 amp. (Thordarson T22R30).

5000 or 10,000 ohms in series with R_1 should do the trick. Or it may be possible to borrow a voltmeter for measuring the output voltage.

The tuning procedure for a regenerative receiver is given earlier in this chapter. Even a short piece of wire hung inside the operating room will serve as an antenna, but for best results an antenna from 30 to 75 feet long, strung as high as possible, should be used.

In buying headphones for use with this receiver, one should avoid the "low-impedance" headphones offered in many of the surplus outlets. While these headsets are excellent when used in the proper circuits, this simple receiver requires the use of "high-impedance" headphones for maximum signal output. Good, inexpensive headphones of this type can be found in any radio store.

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. 5-33, 5-35 and 5-36. As can be seen from the circuit in Fig. 5-34, 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 only 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.

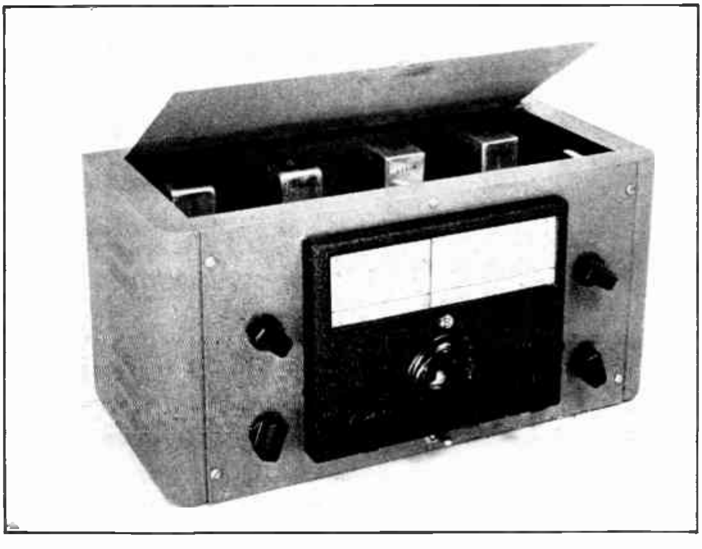


Fig. 5-33 — An amateur-band eight-tube receiver. The knobs on the left control audio volume (upper) and b.f.o. pitch, and the two on the right handle r.f. and i.f. gain (upper) and i.f. regeneration. The knob to the left of the large tuning knob is fastened to the *MAN.-A.V.C.-B.F.O* switch, and the 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.

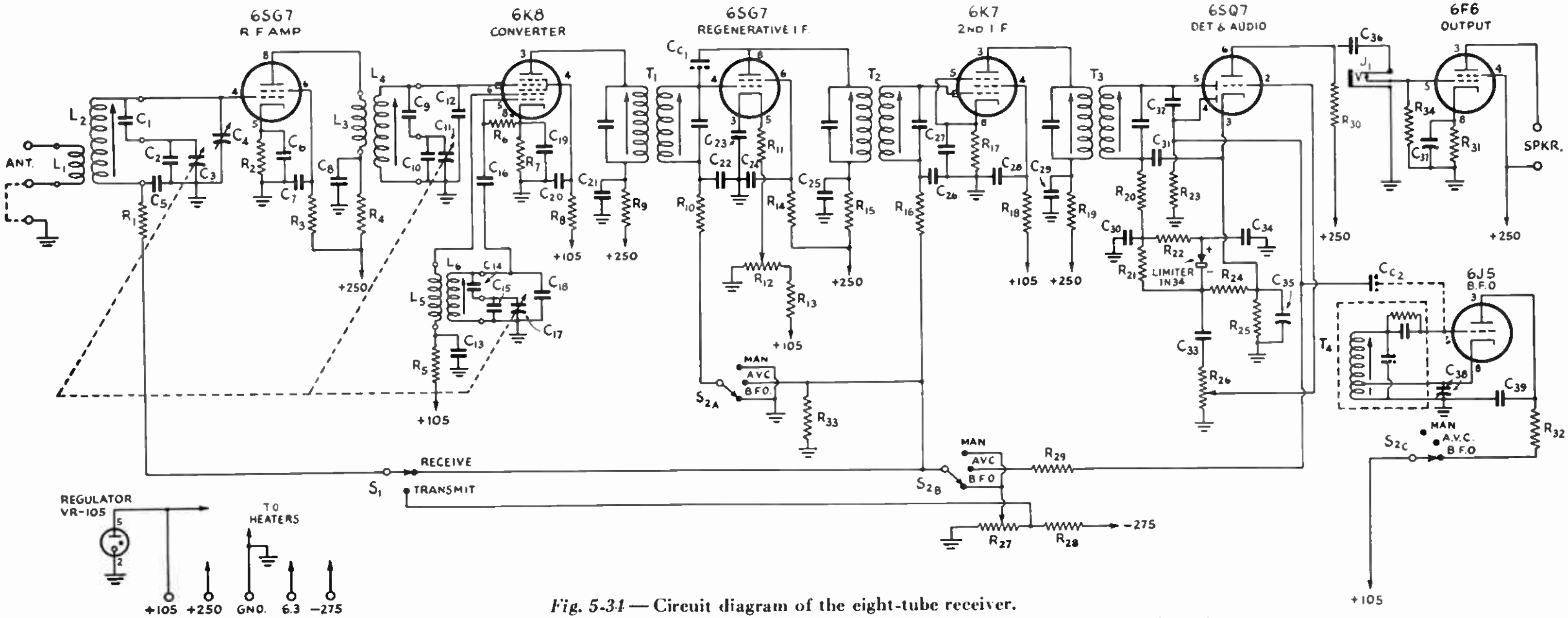


Fig. 5-34 — Circuit diagram of the eight-tube receiver.

C_1, C_9, C_{14} — See coil table.
C_2, C_{10}, C_{12}, C_{18} — 10-μμfd. ceramic.
C_3, C_{11} — 15-μμfd. midget variable (National UM-15).
C_4 — 15-μμfd. 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-μfd. mica or ceramic.
C_{15} — 37-μμfd. ceramic (10 and 27 in parallel).
C_{16}, C_{30}, C_{32} — 100-μμfd. mica.
C_{17} — 35-μμfd. midget variable (National UM-35).
C_{31} — 220-μμfd. mica.
C_{33} — 0.05-μfd. paper, 200 volts.
C_{34} — 0.1-μfd. paper, 200 volts.
C_{35}, C_{37} — 10-μfd. 25-volt electrolytic.
C_{36} — 0.1-μfd. paper, 400 volts.
C_{38} — 35-μμfd. 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).

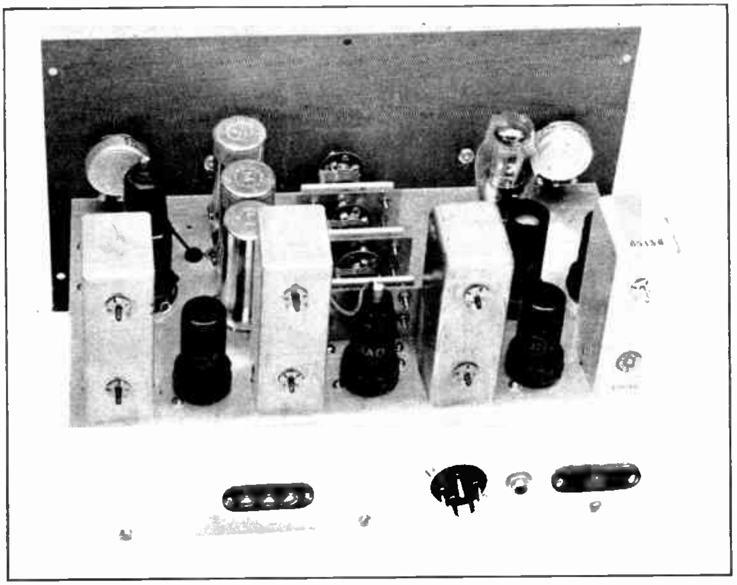


Fig. 5-35 — This view of the eight-tube receiver chassis shows the mounting of the tuning condensers and the placement of most of the large components. The three shielded plug-in coil assemblies can be seen to the left of the tuning gang. The 6K8 converter is the tube on the left nearest the panel.

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 7¼ inches deep on the top. It is 3⅜ inches deep at the rear and ⅛ inch less at the front. The rear edge is reinforced with a piece of ⅜-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 ⅝-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 2½ inches wide and 1 9/16 high, with ½-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 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.

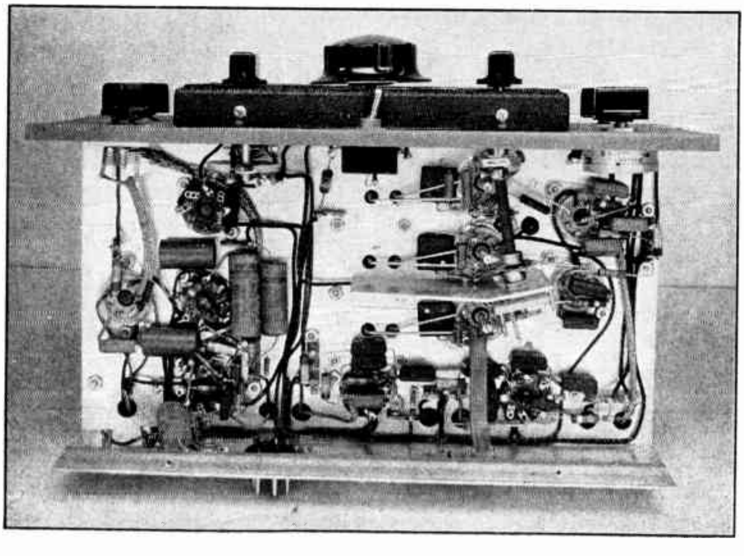
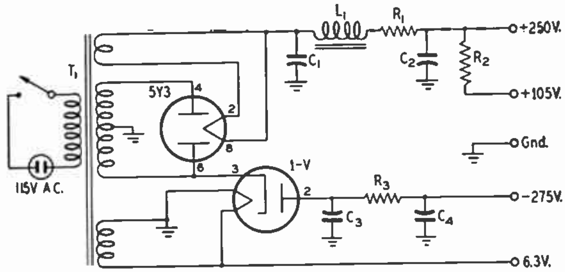


Fig. 5-36 — The mica by-pass condensers used throughout the r.f. and i.f. stages are grouped around the sockets of their respective tubes. Tie-points are used wherever necessary to support small resistors and condensers. The antenna trimmer condenser is mounted on a bracket which also serves as shielding 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.

Fig. 5-37 — Wiring diagram of power supply for the eight-tube receiver.
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.

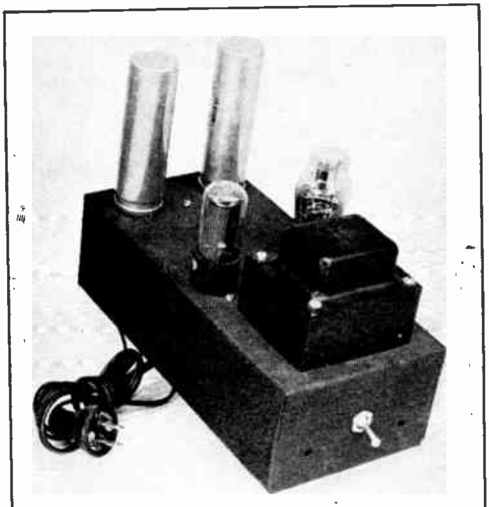


Fig. 5-38 — Power supply for the eight-tube receiver. Two rectifiers are required because a separate supply is incorporated for gain-control purposes. The filter choke and the negative-supply filter condensers are mounted under the chassis. At the rear of the chassis is the socket for the power cable.

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, saving the first tube a little if the transmitter is pouring some power into the receiver.

Power Supply

A power supply suitable for the eight-tube receiver is shown in Figs. 5-37 and 5-38. An idea of the parts arrangement can be obtained from Fig. 5-38, 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.

The filtering of the power supply is quite adequate and no trace of hum should be found in the completed receiver when used with this power supply. If any a.c. hum is noticed, it is being introduced in the audio section if it is still present with the r.f. gain control set at minimum. Probable sources of hum in the audio system are leads to C_{33}, R_{26}, C_{36} or J_1 running too close to a "hot" (ungrounded) heater lead, and the correction is to remove these leads from the field of the heater wiring. If signals are modulated with a.c. hum, particularly at the higher frequencies, it is possible that the grid circuit of the 6K8 converter is picking up hum from a nearby heater lead.

A Simple Audio Noise Limiter

The limiter shown in Fig. 5-39 is plugged into the receiver headphone jack and the headphones are plugged into the limiter, with no work required on the receiver. The limiter will cut down serious noise on 'phone signals, and it will keep the strength of c.w. signals at a constant level. It will do much to relieve the operating fatigue caused by long hours of listening to static crashes, key clicks encountered on the air and with break-in operation, and the like.

Fig. 5-39 — **A simple audio noise limiter for reducing operator fatigue caused by ignition noises, key clicks and static crashes.**

The wiring diagram, Fig. 5-40, shows how two 1N34 crystal diodes are individually biased by 1½-volt flashlight cells. The crystals short circuit any audio signal that has an amplitude of more than 3 volts peak-to-peak. A 10,000-ohm potentiometer, R_2, allows the operator to control the output from the limiter to his headphones and is useful in establishing the optimum relationship between the receiver volume-control setting and the headphone signal strength. A 6AL5 twin diode can be substituted for the two 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 bias cells, and their useful life will be their shelf life.

The limiter can be built in a 4 × 4 × 2-inch cabinet, as shown in Fig. 5-41. The front panel carries the "on-off" switch, the headphone jack and the potentiometer. The 1N34 crystals are mounted on their own leads. Care must be taken while soldering to hold the leads of the crystal diodes with long-nose pliers placed between the point being soldered and the body of the crystal. The pliers conduct away the heat that might otherwise damage the crystal.

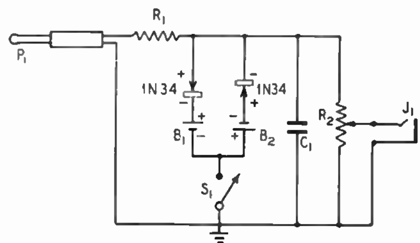


Fig. 5-40 — **Wiring diagram of the audio noise limiter.**
C_1 — 0.0022-μfd. mica.
R_1 — 15,000 ohms, 1 watt.
R_2 — 10,000-ohm potentiometer, wire-wound.
B_1, B_2 — 1½-volt flashlight cell.
J_1 — Open-circuit Jack.
P_1 — Headphone plug.
S_1 — S.p.s.t. toggle switch.

The back panel carries the batteries. A wooden stirrup has contacts of folded copper braid that make contact to one end of the batteries, and a strip of Presdwood with similar contacts is used at the opposite end. The batteries are secured to the panel and the two strips under tension with rubber bands tied to hooks made from soldering lugs.

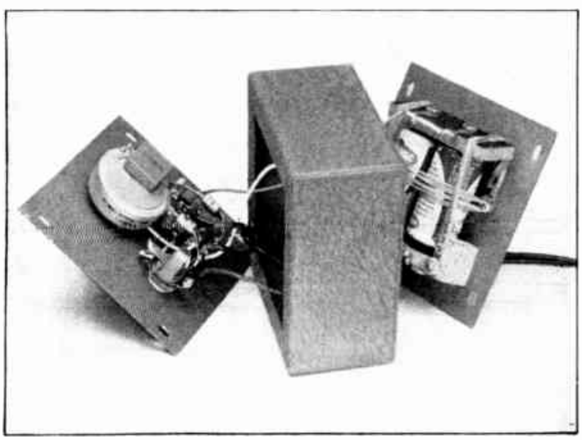


Fig. 5-41 — **The audio noise limiter is built on the two removable panels of a small cabinet. The dry cells are held in place by rubber bands.**

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. 5-42 and 5-43 can be used. The wiring diagram, Fig. 5-44, is an adaptation of Fig. 5-22C, 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 amplifier controlled by the a.v.c. If possible and desirable, the meter and circuit can be built into the receiver.

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."

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.

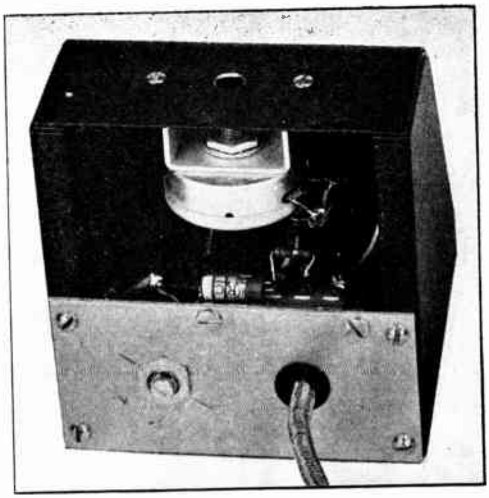


Fig. 5-43 — This rear view of the S-meter shows the meter shunt, R_1, and a tie-point strip mounted on a metal strip attached to the rear side of the meter cabinet. Resistors R_2, R_3 and R_5 are mounted on the tie-point strip. A three-wire cable, running out of the case through a rubber grommet, connects the meter to the receiver.

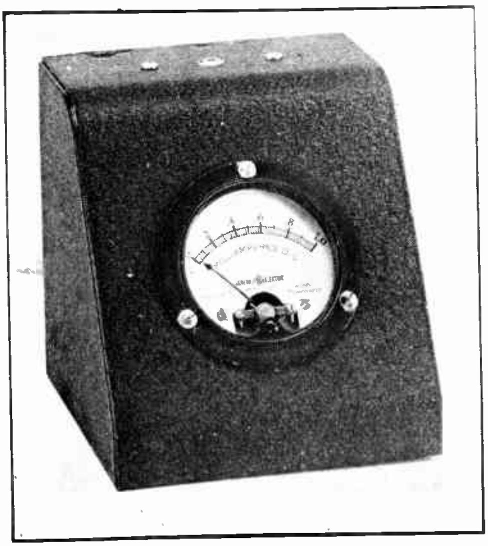


Fig. 5-42 — Front view of the signal-strength indicator. The 0–1 milliammeter is mounted in a metal meter case. The zero-adjustment potentiometer, R_4, is mounted below the top of the cabinet by means of a "U"-shaped bracket; the potentiometer shaft is slotted so that it can be adjusted with a screwdriver. 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.

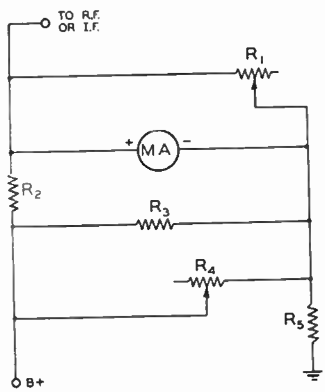


Fig. 5-44 — Wiring diagram of the signal-strength indicator.

R_1 — 100-ohm wire-wound potentiometer.
R_2 — 220 ohms, ½ watt.
R_3 — 680 ohms, ½ watt.
R_4 — 1000-ohm wire-wound potentiometer.
R_5 — 47,000 ohms, 1 watt.
MA — 0–1 ma. d.c. meter.

A Peaked Audio Amplifier

The peaked audio amplifier shown in Figs. 5-45 and 5-47 uses only resistors and condensers to obtain a high degree of selectivity. The circuit, Fig. 5-46, consists of an ordinary audio amplifier and a simple twin-"T" resistance-capacitance bridge. The bridge has a null at the desired audio frequency, and the bridge is connected in a negative-feed-back loop in the amplifier. As a result, the amplifier is highly degenerative at all frequencies except that at which the bridge circuit shows a null. By controlling the amount of negative feed-back, varying degrees of selectivity can be obtained.

The unit, minus its power supply, is housed in a 3 × 4 × 5-inch standard steel box. To simplify construction, most of the components are mounted on a piece of 4 × 5 × 1/16-inch aluminum that replaces one of the removable panels of the box.

When completed and connected to a source of plate and heater power — the plate demand is about 20 ma. at 250 volts — plug P_1 into the receiver output jack and the headphones into J_1. Set the selectivity control, R_8, at maximum, i.e., with the arm farthest away from the grounded end. Tune in a stable c.w. signal and adjust C_8 until the amplifier "rings" or indicates a tendency toward oscillation. Back off on R_8 until you can tune through a peak on C_8 without oscillation, and the audio amplifier is adjusted. In operation, the control for R_8 can be advanced or backed off to give the desired amount of selectivity.

If the amplifier is used with a single-signal superheterodyne in which the crystal filter already contributes considerable selectivity, it is essential that the b.f.o. be adjusted to give peak audio response from the receiver at the frequency for which the audio amplifier shows maximum gain.

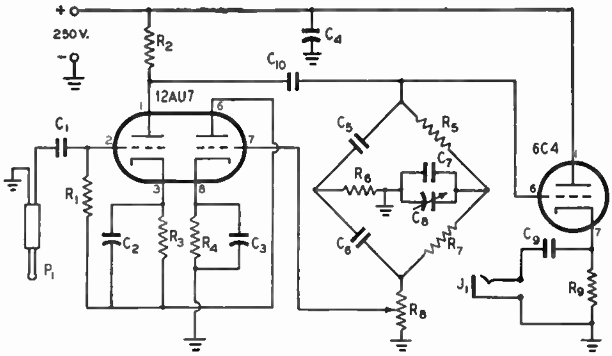


Fig. 5-46 — Wiring diagram of the peaked audio amplifier.

C_1, C_{10} — 0.01-μfd. paper.
C_2, C_3 — 25-μfd. 25-volt electrolytic.
C_4 — 8-μfd. 450-volt electrolytic.
C_5, C_6 — 680-μμfd. mica.
C_7 — 0.001-μfd. mica.
C_8 — 280–1050-μμfd. mica compression trimmer (El-Menco 306).
C_9 — 0.1-μfd. 200-volt paper.
R_1 — 1.0 megohm.
R_2 — 56,000 ohms, 1 watt.
R_3, R_4 — 1200 ohms.
R_5, R_7 — 0.22 megohm.
R_6 — 0.1 megohm.
R_8 — 2.0-megohm volume control.
R_9 — 10,000 ohms, 1 watt.
Resistors are ½-watt composition unless specified otherwise.
J_1 — Open-circuit jack.

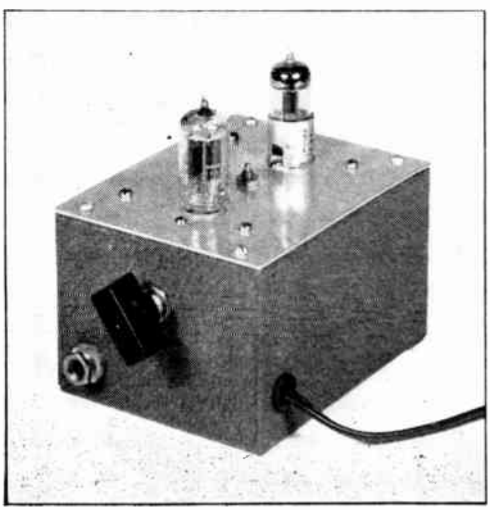


Fig. 5-45 — A peaked audio amplifier for increased c.w. selectivity. It is connected to the receiver at the headphone jack, and the headphones plug into the unit. The knob controls the degree of selectivity.

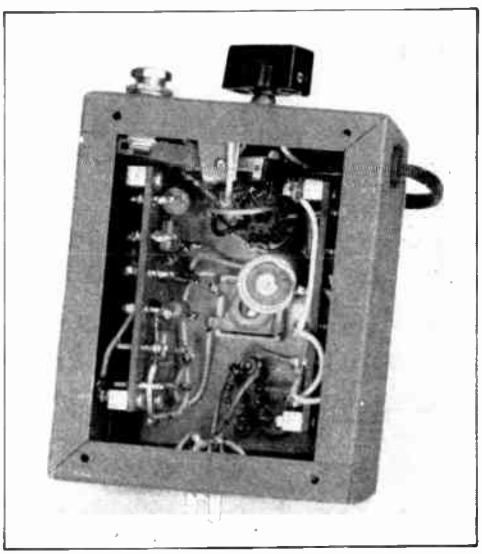


Fig. 5-47 — Construction of the peaked audio amplifier is facilitated by mounting the parts on an aluminum panel that replaces the normal panel of the cabinet. Two resistor boards, supported by square posts mounted on the panel, are used to support most of the small components. A Jones P-304-AB base-mounting plug on the cabinet is used for connecting to the power supply.

A Bandswitching Preselector for 14 to 30 Mc.

The performance of many receivers begins to drop off at 14 and 30 Mc. The signal-to-noise ratio is reduced, and trouble with r.f.-image signals becomes apparent. The preselector shown in Figs. 5-48 and 5-50 can be added ahead of any receiver without making any changes within the receiver, and a self-contained power supply eliminates the problem of furnishing heater and plate power.

As can be seen from the wiring diagram, Fig. 5-49, a 6AK5 r.f. pentode is used in the preselector. Both the grid and plate circuits are tuned, but the tuning condensers are ganged and only one control is required. The gain through the amplifier is controlled by changing the cathode voltage, through R_3. A selenium rectifier is used to supply plate power, and the heater power comes from a step-down transformer. The chassis is at r.f. ground but the d.c. circuit is isolated, to prevent short-circuiting the a.c. line through external connections to the preselector.

A two-section ceramic switch selects either the 14- to 21-Mc. or the 28-Mc. coil, or the antenna can be fed through directly to the receiver input. When operating in an amateur band between 14 and 30 Mc., switching to the band not in use will attenuate one's own signal sufficiently to permit direct monitoring, in most cases.

As shown in Fig. 5-48, the ganged condensers are controlled from the front panel by a National MCN dial, and a small knob to the right of this dial is connected to the antenna trimmer, C_4, for peaking the tuning with various antennas. The a.c. line is controlled by S_2, a toggle switch mounted on the panel.

The preselector is built on a 3 × 5 × 10-inch chassis, and a 6 × 6-inch plate of thin metal is used for a panel. A 1¾ × 3-inch aluminum bracket mounted about 3½ inches behind the front panel supports the tuning condenser, C_5, and the antenna trimmer, C_4. Millen 39005 flexible couplings are required to handle the offset shaft of C_4. Both C_5 and C_8 are mounted on the chassis with 6-32 screws, but the chassis should be scraped free of paint before installation, to insure good contact.

The shield partition between the two switch sections (Fig. 5-50) straddles the tube socket and shields the grid from the plate circuit. The switched ends of all coils are supported by their respective switch points, and the other ends are soldered to tie points mounted on the chassis. The mica trimmers, C_9 and C_{10}, are supported on short lengths of stiff wire, and a hole in the side of the chassis is required to reach C_{10} with an aligning tool.

The power-supply components are mounted as near the rear of the chassis as possible. The selenium rectifier must be insulated from the chassis.

COIL TABLE FOR THE PRESELECTOR

L_1	5 t. No. 24, ¾-inch diameter (B & W 3012)
L_2	5 t. No. 24, 1-inch diameter (B & W 3016)
L_3	6 t. No. 24, ¾-inch diameter (B & W 3012)
L_4	7 t. No. 20, 1-inch diameter (B & W 3014)
L_5	7½ t. No. 20, ¾-inch diameter (B & W 3010)
L_6	3 t. No. 24, 1-inch diameter (B & W 3015)
L_7	11 t. No. 24 d.c.c., close-wound, ½-inch diameter
L_8	4 t. No. 28 d.c.c., close-wound, ½-inch diameter

L_7 and L_8 are wound adjacent on a ½-inch diameter polystyrene form (National PRD-2)

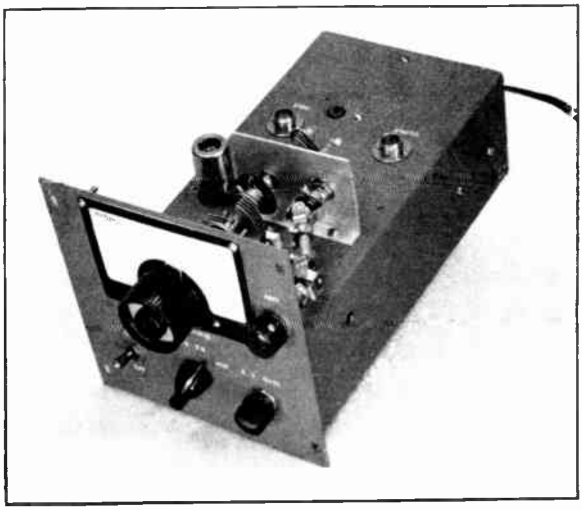


Fig. 5-48 — A bandswitching preselector for 14 and 28 Mc. A single 6AK5 amplifier is used, and the power supply is included in the unit. The antenna-trimming condenser is mounted on the small aluminum partition.

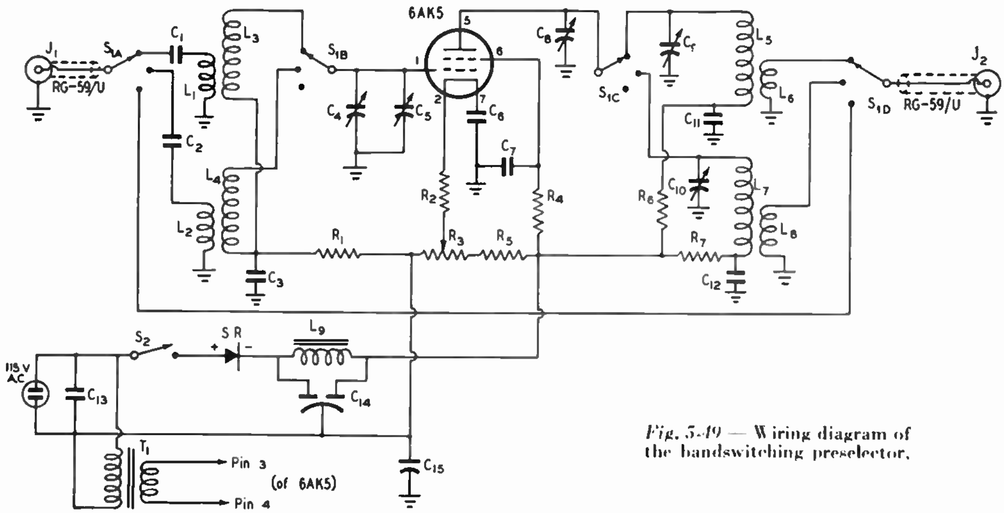


Fig. 5-49 — Wiring diagram of the bandswitching preselector.

C_1, C_2 — 10-μμfd. mica.
C_3, C_6, C_7, C_{11}, C_{12} — 680-μμfd. mica.
C_4 — 15-μμfd. midget variable (Millen 20015).
C_5, C_8 — 50-μμfd. midget variable (Millen 19050).
C_9, C_{10} — 3- to 30-μμfd. mica trimmer.
C_{13}, C_{15} — 0.01-μfd. paper, 400 volts.
C_{14} — Dual 40-μfd. 150-volt electrolytic.
R_1 — 27,000 ohms.
R_2 — 330 ohms.
R_3 — 5000-ohm wire-wound potentiometer.
R_4 — 1700 ohms.
R_5 — 18,000 ohms, 2 watts.
R_6, R_7 — 470 ohms.
L_1–L_8 — See coil table.
L_9 — 20-henry 30-ma. filter choke.
J_1, J_2 — Coaxial-cable jack (Jones S-401).
S_1 — 2-gang 2-circuit 5-position ceramic (Mallory 177C).
S_2 — S.p.s.t. toggle.
SR — 50 ma. selenium rectifier.
T_1 — 6.3-volt transformer.

The coils are made from B & W "Miniductors," as shown in the coil table, with the exception of one plate and coupling coil which are wound on a polystyrene form. The ground returns for the cathode and plate by-pass condensers are made to a common terminal, a soldering lug under one of the mounting screws for C_8.

When the wiring has been completed and checked, the antenna is connected to J_1 and a cable from J_2 is run to the receiver input. Tune the receiver to the 14-Mc. band and set S_1 to the proper point. Then turn the main tuning dial until the noise or signal increases to a maximum. This should occur with C_5 and C_8 set at close to maximum capacity. Then peak the noise by adjusting C_{10} and C_4.

The 28-Mc. range is adjusted in the same way, with the exception that C_9 is touched up. It may be found necessary to touch up C_4 when different antennas are used. The preselector may oscillate with no antenna connected, but with any type of wire or feedline the operation of the amplifier should ordinarily be perfectly stable.

As shown, the preselector is intended for use with coaxial-line feed to the antenna and to the receiver. If a balanced two-wire line is used from the antenna, it is recommended that a suitable two-wire connector be substituted for J_1. The grounded sides of L_1 and L_2 should be disconnected from ground and returned to one side of the connector. The output connector can be left as shown, since at the lower frequencies the proper antenna connection isn't so important.

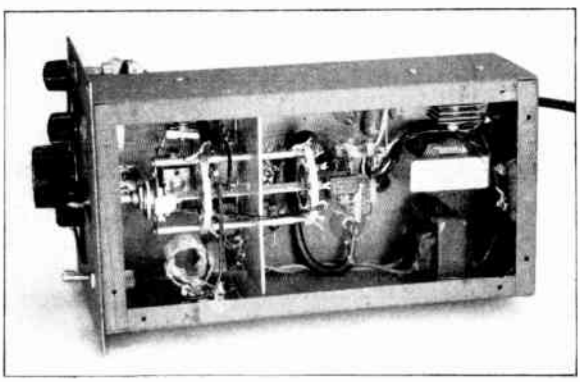


Fig. 5-50 — A view underneath the chassis of the bandswitching preselector, showing the shield partition between switch sections and the selenium rectifier and associated filter.

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. One way to do this is to use, in reverse, any of the line-coupling devices advocated for use with a transmitter. Naturally the components can be small, because the power involved is negligible, and small receiving condensers and coils are quite satisfactory. Some provision for adjustable coupling is recommended, as in the transmitting case, because the signal-to-noise ratio at 14 and 28 Mc. is dependent, to a large extent, on the degree of coupling to the antenna system. The tuning unit can be built on a small chassis located near the receiver, or it can be mounted on the wall and a piece of RG-59/U run from the unit to the receiver input, in the manner of a link line in transmitting practice. For ease in changing bands, the coils can be switched or plugged into a suitable socket. Adjustable coupling not only offers an opportunity to adjust for best signal-to-noise ratio, but the coupling can be decreased when a strong local signal is on the air, to eliminate "blocking" and cross-modulation effects in the receiver.

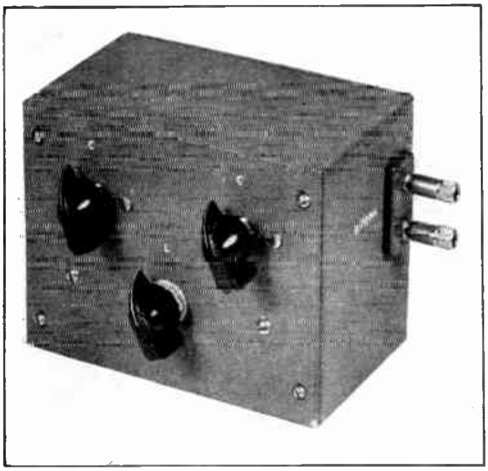


Fig. 5-51 — A compact coupling network for matching a balanced line to the receiver on 14 and 28 Mc.

One convenient type of antenna-coupling unit for receivers uses the familiar pi-section filter circuit, and can be used to match a wide range of antenna impedances. The diagram of a compact unit of this type is shown in Fig. 5-52. Through proper selection of condensers and inductances, a match can be obtained over a wide range of values. The device 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 antenna 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 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, 11½ turns and 30 turns. All of the wiring, with the exception of the leads to the input and output terminals, can be done with the panel removed from the box.

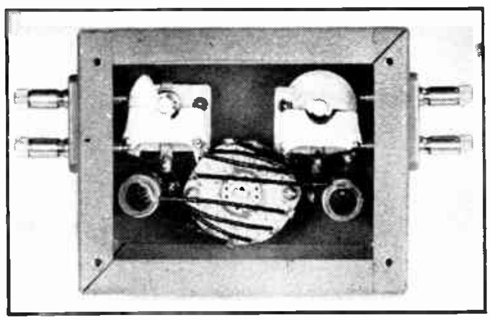


Fig. 5-53 — Rear view of the antenna-coupling unit. The two coils can be seen directly below the two tuning condensers.

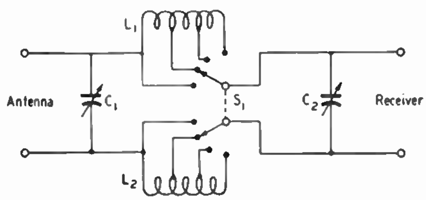


Fig. 5-52 — Circuit diagram of the coupling unit.

C_1, C_2 — 100-μμfd. midget variable (Millen 22100).
L_1, L_2 — 30 turns No. 24 d.c.c. close-wound on ½-inch diameter polystyrene form, tapped at 2½, 6½ and 11½ turns.
S_1 — 2-circuit 5-position single-section ceramic wafer switch (Mallory 173C).

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. The small capacity remaining has a negligible effect.

Receiver Matching to Coaxial Line

While some of the war-surplus receivers are designed to work from a low-impedance antenna, most of the popular communications receivers are designed for an impedance of from 300 to 500 ohms. When using coaxial-line feed from an antenna, as is not rare on 14 and 28 Mc., maximum signal transfer from line to receiver is not obtained unless some type of matching network is used. The pi-section coupler can be used, by short-circuiting the inductance in one leg and connecting this side of the coupler to the outer conductor of the cable and to the ground connection on the receiver. However, in matching between two unbalanced resistive loads of this type, another and slightly simpler circuit can be used. It is called an "L" section.

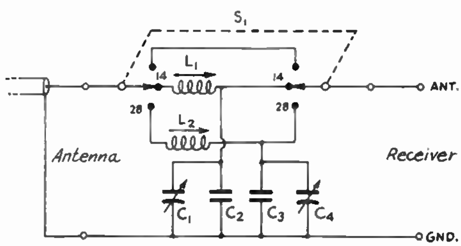


Fig. 5-54 — Wiring diagram of the "L"-section matching network.

C_1, C_4 — 3- to 30-μμfd. mica-compression trimmer.
C_2 — 100-μμfd. mica.
C_3 — 47-μμfd. mica.
L_1 — 12 turns, spaced to occupy 5/8 inch.
L_2 — 7 turns, spaced to occupy 7/16 inch. L_1 and L_2 wound with No. 18 d.s.c. on National XR-50 (½-inch diameter) iron slug-tuned forms.
S_1 — 2-pole 3-position rotary wafer switch.

An "L"-section matching coupler for 14 and 28 Mc. is shown in Fig. 5-55. All of the components are mounted on a switch, and the unit is intended to be mounted on the antenna and ground post of the receiver. As can be seen from the wiring diagram in Fig. 5-54,

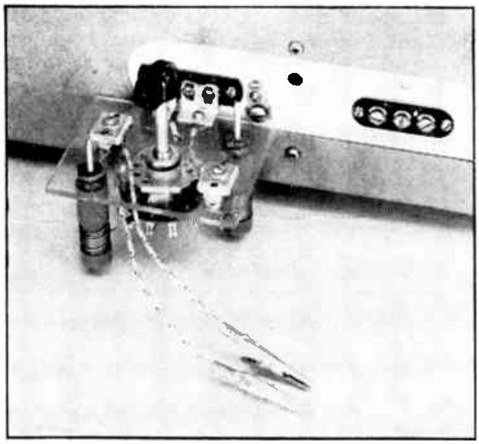


Fig. 5-56 — The "L"-section coupler mounted on the antenna and ground binding posts of a communications receiver.

provision is included for straight-through operation between feed line and receiver on the other frequencies.

The values of the components are not critical, but provision is included for adjusting both the inductance and the capacity, to accommodate minor variations in receiver impedances. If operation is limited to one band, or if different receivers or converters are used on the various bands, the coil and condenser can be mounted right at the receiver terminals without the switch. As shown, the unit is intended for use following an antenna change-over relay, and it is assumed that the different antennas are changed at the relay. However, if a relay is not used, the different feed lines can be brought directly to the unit and soldered to the antenna sides of L_1 and L_2.

The units can be adjusted on a local signal that is not fading, by adjusting the inductance and capacity for maximum signal, as indicated by the S-meter. It is not be to expected that the adjustment will be critical, but the gain obtained by proper matching will be observed by switching to the straight-through condition, and comparing the difference. The improvement will be only slight if the initial mismatch was small, but an improvement of several db. can be expected in any case.

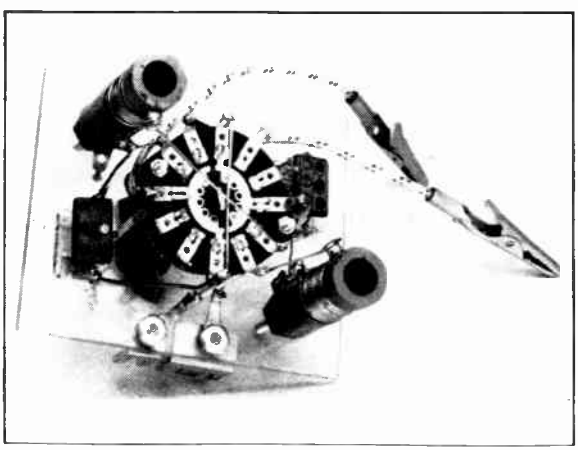


Fig. 5-55 — An "L"-section matching network for coupling the receiver to coaxial line. It is designed for use between 50- or 75-ohm line and a receiver of 300 to 400 ohms input impedance.

Receiver Matching to Tuned Lines

The pi-section coupler shown in Figs. 5-51, 5-52 and 5-53 can be used in many instances for matching a balanced open-wire line to the receiver, and it can be used with an unbalanced line by short-circuiting the inductance in the grounded side of the unbalanced line. However, there are many applications where another type of coupler is slightly more advantageous, as when an all-band antenna system with tuned feeders is used, or where a wide range of line impedances may be encountered. This other type of coupler, shown in Figs. 5-57, 5-58 and 5-59, is simply a scaled-down transmitter coupler, with provision for either series or parallel tuning. The change from series to parallel tuning is made simply by the manner in which the antenna connection plate is plugged into the unit.

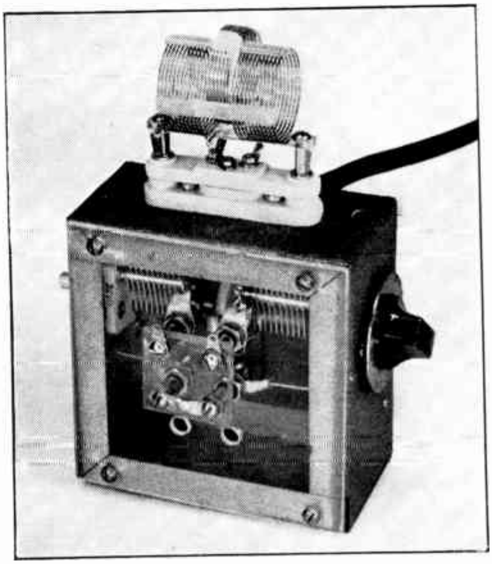


Fig. 5-57 — **A small tuned coupler for matching the receiver to a tuned line. The unit is made either series- or parallel-tuned by the position of the antenna connection block.**

As can be seen in the wiring diagram, Fig. 5-58, when the antenna connection plate is plugged in so that all four contacts are engaged, the two condensers are connected across the coil in series, to give parallel tuning. When the plate is dropped down, so that only the antenna plugs engage at *A* and *B*, the unit is connected for series tuning. Small low-power transmitting coils with swinging links are used.

The unit is built in a 4 × 4 × 2-inch box, with the coil socket mounted on one 2 × 4-inch side. One of the 4 × 4-inch side plates is replaced by a sheet of polystyrene or other insulating material, on which are mounted four banana jacks. A similar but smaller piece of insulating material is drilled at the same time to take four banana plugs. A pair of clearance holes must be added to the larger plate to clear two of the plugs when the series connection is used.

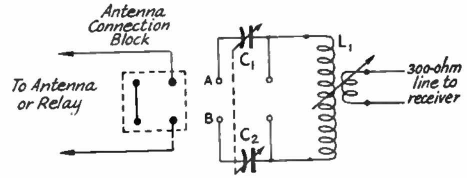


Fig. 5-58 — **Circuit of the tuned antenna coupler.**
C_1, C_2 — 100-μμfd. midget variable (Millen 22100).
L_1 — Coil to tune to band in use, with swinging link (National AR-16).

The two condensers are mounted in the box and ganged with an insulated shaft coupling. The remaining 4 × 4-inch side plate is drilled and filed to form an oval hole that will pass the 300-ohm line from the coupler to the receiver. A rubber grommet should be fitted in the hole to protect the line from the metal and to provide a little clearance.

In operation, the coupler is used in exactly the same way that one is used with a transmitter. Some experimenting is necessary to determine whether series or parallel tuning should be used on the various bands, and it may be necessary to use the coil from the next lower-frequency band if series tuning is indicated, or to remove a few turns from a coil if parallel tuning is required. In any event, the tuner should tune fairly sharply and give a definite "peak" to the incoming signals. When this condition has been found on any one band, the coupling can then be adjusted for maximum response to the signals, by adjusting the position of the link winding within L_1.

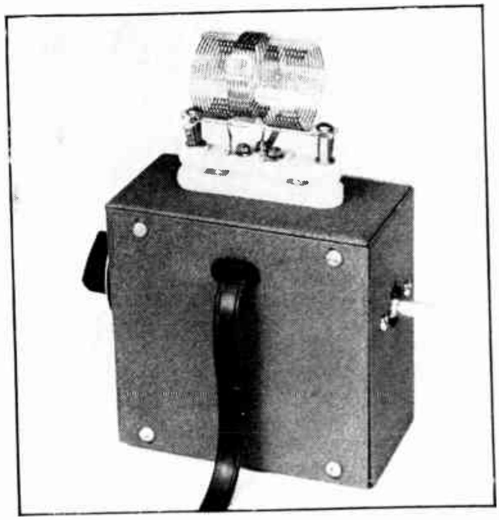


Fig. 5-59 — **Another view of the tuned antenna coupler.**

A One-Tube Converter for 10 and 11 Meters

The 10- and 11-meter converter shown in Figs. 5-60 and 5-62 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 bandspread 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. 5-61, 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.

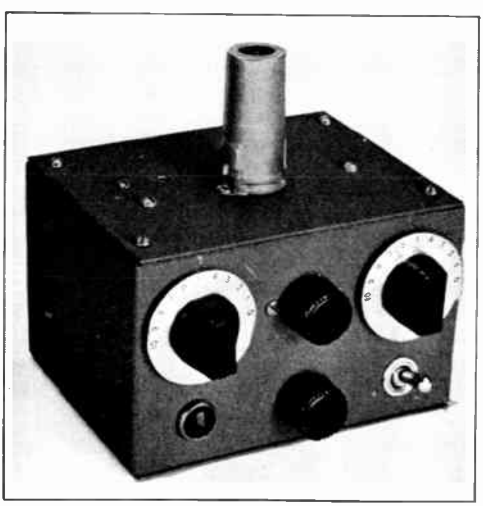


Fig. 5-60 — A front view of the ten-meter converter. The components and controls on the front wall of the case, from left to right, are as follows: top row, r.f. tuning control, oscillator tuning knob, and i.f.-circuit control; bottom row, dial-light assembly, antenna change-over switch, and filament 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. 5-62. 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 hookup 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.

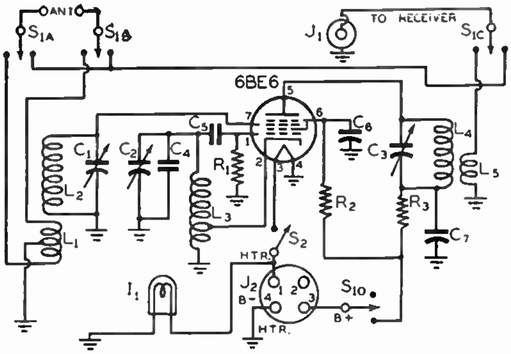


Fig. 5-61 — Circuit diagram of the low-cost ten-meter converter.

C_1, C_2 — 15-μμfd. variable (Millen 20015).
C_3 — 75-μμfd. variable (Millen 20075).
C_4 — 100-μμfd. silver mica.
C_5 — 47-μμfd. mica.
C_6, C_7 — 0.01-μfd. 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 turns 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.

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. to supply the plate and screen power.

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 in the recommended range of 90 to 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 limits of 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, but they require little or no adjusting in the 11-meter band.

It is important that the link between the converter be well shielded, to avoid picking up any signals in the tuning range of the receiver. A length of RG-58/U or RG-59/U should be used between the converter and the receiver and, if necessary, a small shield should be mounted over the antenna binding post on the receiver. If it is found to be impossible to keep out some particularly strong local signal that is being picked up on the coupling lead, it may be necessary to shift the tuning range of the receiver (by resetting C_2) to avoid this signal. Such a condition is very unusual, however, if care is taken with the coupling lead.

If no communications receiver is available, a war-surplus BC-454 aircraft receiver (tuning range of 3 to 6 Mc.) makes an inexpensive receiver for use with this converter.

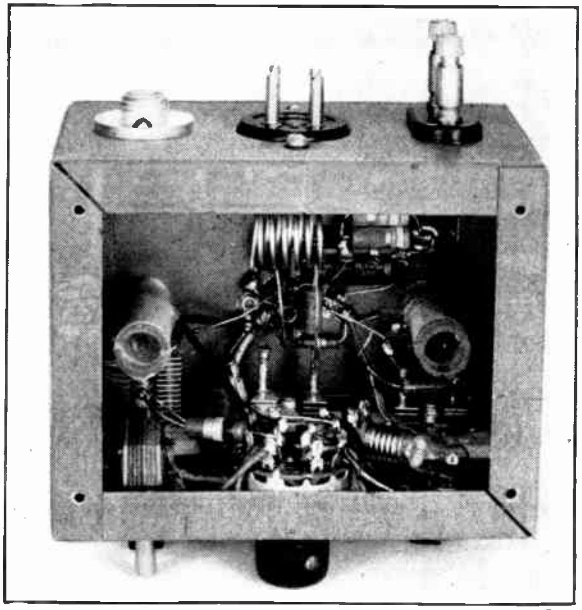


Fig. 5-62 — An inside view of the ten-meter converter. The r.f. and i.f. coils are at the right and left ends of the box, respectively. The oscillator coil may be seen to the rear of the tube socket. This view also shows the arrangement of the components mounted on the front wall of the case and the 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.

Crystal-Controlled Converters for 14, 21 and 28 Mc.

The principle of using a fixed high-frequency oscillator in a converter and tuning the receiver the converter works into can be elaborated upon by using a stage of r.f. amplification ahead of the mixer and by using a crystal-controlled oscillator for maximum stability. Since such a converter is generally used on a high frequency where fundamental crystals are not available, it is necessary to use a harmonic of a lower-frequency crystal. A crystal-controlled converter of this type is shown in Figs. 5-63 and 5-65. A separate converter is required for the 14-, 21- and 27-/28-Mc. bands, since by using separate converters it is possible to simplify their construction and to maximize their performance.

The converter uses the harmonic of a crystal oscillator to provide an exceedingly stable high-frequency oscillator signal. For example, in the 10-meter converter a 12.25-Mc. crystal doubles to 24.5 Mc., and this signal is fed to the mixer. By tuning the amplifier (your present receiver) following the mixer over the range 3.5 to 5.2 Mc., you are, in effect, tuning across the 28-Mc. band. The r.f. circuits in the converter are tuned to 28 Mc., and only have to be touched up when going from one end of the band to the other.

The wiring diagram is shown in Fig. 5-64. A neutralized triode-connected 6AK5 is used for the r.f. amplifier. There is some question as to its necessity on 14 and 21 Mc., where the atmospheric noise is generally high enough to limit the maximum usable sensitivity. A pentode-connected 6AK5 could probably be used with no detectable difference in performance on 14 and 21, but the triode is easy to handle and you don't lose anything by using it. Using high-impedance circuits with the pentode might give trouble from regeneration, unless the stage were neutralized. Adjustable antenna coupling and a Faraday screen are included to accommodate various antenna systems and to eliminate capacity coupling to the antenna line. The r.f. stage runs at 105 volts on the plate, since this gives the best noise figure. The separate plate lead also offers an opportunity to kill the converter by opening this circuit. The 6AK5 pentode mixer is easy to handle and quiet enough so that its noise doesn't impair the over-all performance. A triode mixer might be used, but the pentode runs with low current and is quiet.

The plate circuit of the mixer is tuned to the center of the receiver tuning range by setting L_4 to resonate with the various shunt circuit capacities. The circuit has a low Q and there is little variation in gain over the range. A 6C4 cathode follower is used as a low-impedance coupling to the receiver input.

One section of a 6J6 twin triode is used for the crystal oscillator, and the other half serves as a frequency multiplier. To minimize the other harmonics existing in the plate circuit of the multiplier, the plate is tapped down on L_6.

To get the best possible r.f. circuits, within the space limitations, B & W "Miniductors" are used for L_1, L_2 and L_3. Their Q is well above that obtainable with smaller-diameter coils, and they are easy to handle. To insure good shielding and low-resistance ground paths, an aluminum chassis is used in preference to the more common steel units.

The converter is built on a 5 × 9½ × 3-inch aluminum chassis, with several shield partitions to reduce unwanted interstage coupling. The most important shield is the one that straddles the r.f. amplifier socket and separates the grid and plate circuits of this stage. The grid tuning condenser, C_2, is mounted on bakelite insulating washers, and its ground lead returns to the common ground at the tube socket, to eliminate stray coupling through chassis cur-

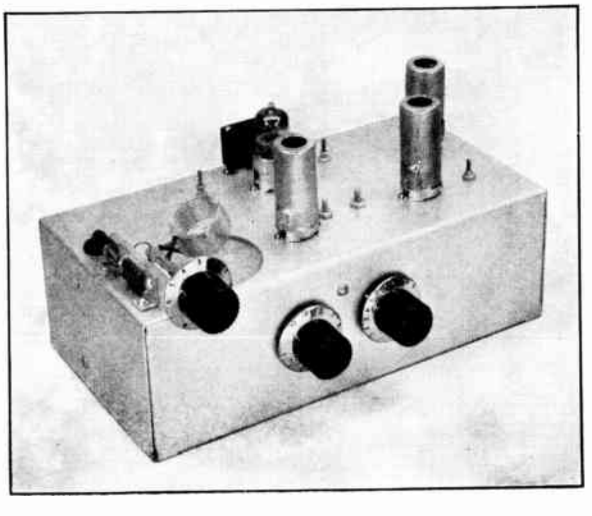


Fig. 5-63 — A 28-Mc. crystal-controlled converter. The adjustable antenna coupling can be seen at the left front. The tube shields, from left to right, cover the triode-connected 6AK5 r.f. amplifier, the 6AK5 mixer and the 6C4 cathode follower. The unshielded tube is the 6J6 oscillator-multiplier.

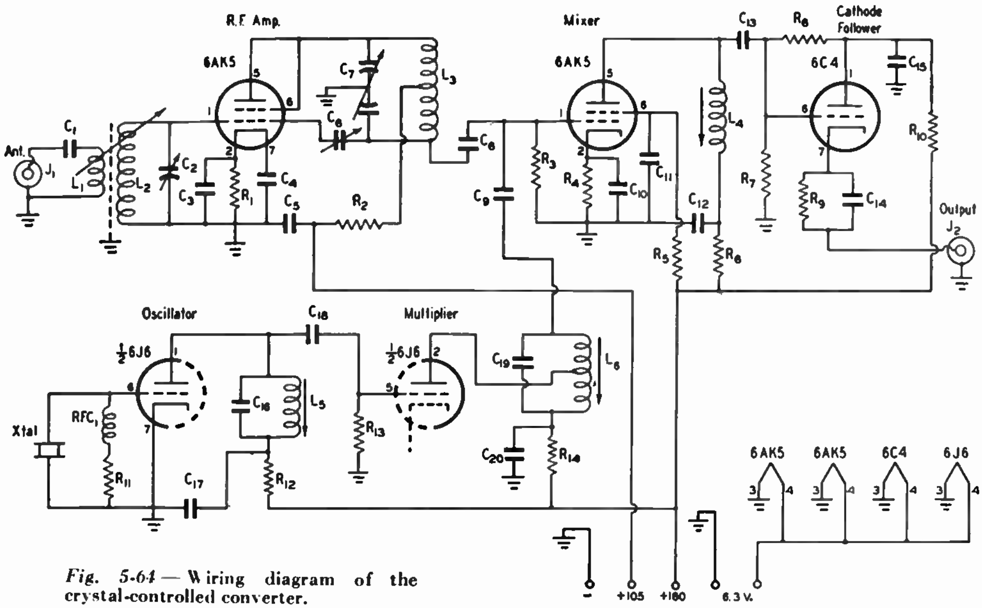


Fig. 5-64 — Wiring diagram of the crystal-controlled converter.

C_1 — 10-μμfd. mica.
C_2 — 20-μμfd. midget variable (Johnson 160-110).
C_3, C_4, C_5, C_{10}, C_{11}, C_{12}, C_{14}, C_{15}, C_{17}, C_{20} — 680-μμfd. mica.
C_6 — 5-μμfd. midget variable (Johnson 160-102).
C_7 — 11-μμfd. midget butterfly (Johnson 160-211).
C_8, C_{13} — 470-μμfd. mica.
C_9 — Twisted wire. See text
C_{16}, C_{19} — See coil table.
C_{18} — 47-μμfd. mica.
R_1, R_9 — 220 ohms.
R_2 — 2200 ohms, 1 watt.
R_3 — 56,000 ohms.
R_4 — 6800 ohms.
R_5 — 0.1 megohm.
R_6, R_{10}, R_{12}, R_{14} — 470 ohms.
R_7, R_{11} — 4700 ohms.
R_8 — 0.18 megohm.
R_{13} — 82,000 ohms.
All resistors ½-watt unless otherwise specified.
L_1, L_2, L_3, L_4, L_5, L_6 — See coil table.
J_1, J_2 — Cable-connector sockets (Jones S-101).
RFC_1 — 750-μh. r.f. choke (National R-33).
XTAL — See coil table.

rents. If this isn't done, you may have trouble neutralizing the amplifier.

A 2¼-inch diameter hole is punched in the chassis, so that the externally-mounted antenna coil, L_1, can be coupled to the grid coil, L_2. The Faraday screen is then mounted across this hole on the underside of the chassis. To construct the Faraday shield, first cut a piece of ⅛-inch-thick polstyrene (Millen Quartz-Q) to measure 2½ by 3¼ inches, and drill a pair of holes at one end to clear No. 6 screws, for mounting the finished shield. (These are the same screws that hold the mounting strip for the antenna condenser, C_1, visible in Fig. 5-63.) At the opposite end of the poly sheet, drill a small hole in each corner, for securing the wire used in making the shield. Then wind No. 20 tinned wire tightly around the poly sheet in the long direction, spacing it with string or more No. 20 wire. When the winding is finished and secured at both ends, unwind the spacing string (or wire) and remove it. If you have done the job carefully, you will have neat parallel lines of wire across the polystyrene, all equally spaced and all lying fairly flat. Then apply two or three heavy coats of Duco cement to *one side only*, allowing sufficient time between coats for the cement to harden thoroughly. When this has been done, it will be found an easy job to cut each wire on the uncemented side. Straighten out the wires so that you now have a flat sheet of parallel wires, and trim off the wires at the mounting holes end of the sheet along a line inside the mounting holes. Figs. 5-65 and 5-66 show what this looks like. When trimming these wires, be careful to see that no wire is left touching an adjacent one. Trim the wire ends at the other end to about ½ inch from the polystyrene. Clamp the shield in a vise, between two pieces of wood, and wrap each wire end around a piece of No. 12 tinned copper, as shown in Fig. 5-66. With a good hot iron, run a bead of solder along the bus, and your shield is finished. Work fast, and no heat will reach the poly. The shield is mounted with the smooth side exposed through the hole, and one end of the No. 12 bus is grounded at the r.f. tube socket.

The grid coil, L_2, is supported by its leads and a couple of drops of Duco cement that hold its grounded end to the Faraday shield. The antenna coil, L_1, is mounted by its leads on a piece of ¼-inch diameter polystyrene rod. The rod is supported by a shaft bushing. A small wire pin through the rod at the back of the bushing and a rubber grommet between the bushing and the control knob give a soft friction lock that holds the coupling in any position. Flexible leads run from the coil to C_1 and the shield of the RG-59/U coaxial line.

The r.f. plate coil, L_3, is cemented to a small piece of polystyrene sheet that is supported by two small brackets. The neutralizing condenser, C_6, is supported by one terminal of C_7 and a stiff wire lead back to the grid pin on the tube socket. The coupling condenser, C_9, is simply an insulated wire wrapped once around the lead from C_8 to the grid of the mixer. It is brought out of the oscillator compartment through a polystyrene or rubber grommet.

After the usual last check of the wiring, connect a power supply and remove the 6AK5 r.f. amplifier from its socket. Listen in on your receiver at the crystal frequency, and if you don't find the crystal signal, adjust L_5 until you do. Then set your receiver on the proper harmonic frequency and peak L_6 for maximum signal, as indicated by your S-meter. When you have done this, you can probably squeeze out a little more by readjustment of L_5. Then back off on L_5 a little, because there is no need to run the crystal at maximum.

COIL TABLE FOR THE CRYSTAL-CONTROLLED CONVERTER

	14 Mc.	*21 Mc.*	*28 Mc.*
L_1	23 t. No. 24 ¾-inch diam. (B & W 3012)	9 t. No. 24 1-inch diam. (B & W 3016)	10 t. No. 20 1-inch diam. (B & W 3015)
L_2	21 t. No. 24 ¾-inch diam. (B & W 3012)	10 t. No. 20 1-inch diam. (B & W 3015)	9 t. No. 20 1-inch diam. (B & W 3015)
L_3	38 t. No. 24 ¾-inch diam., center-tapped (B & W 3012)	22 t. No. 24 ¾-inch diam., center-tapped (B & W 3012)	16 t. No. 24 ¾-inch diam., center-tapped (B & W 3012)
L_4	Slug-tuned coil (Cambridge Thermionic Corp. 1-Mc. LSM with 200 turns removed) (Coils for L_5 and L_6 are wound on ¼-inch diameter Cambridge Thermionic Corp. LSM forms)		
L_5	No. 32 enam., close-wound, ½ inch long	No. 32 enam., close-wound, ½ inch long	30 t. No. 28 enam., close-wound
L_6	22 turns No. 28 enam., close-wound, center-tapped	20 t. No. 20 enam., close-wound, center-tapped	20 t. No. 24 enam., close-wound, center-tapped
C_{16}	75 μμfd.	75 μμfd.	33 μμfd.
C_{19}	0	22 μμfd.	22 μμfd.
Xtal	6000 kc. (triples)	5875 kc. (triples)	12,250 kc. (doubles)

Then tune your receiver — its antenna circuit must complete the cathode circuit of the 6C4 follower — to about 3.8 Mc. and peak L_4 for maximum noise. The adjustment is not sharp, because of the low Q of the circuit. If your receiver has an antenna trimmer, don't forget to peak it, too. Then plug in the 6AK5 r.f. amplifier and, after the tube has warmed up, rock C_2 and C_7. Unless you are very lucky, you will find several settings where you are greeted by birdies and squawks. Through the hole in the bottom plate, use an alignment tool to adjust C_6 a little at a time, until you lose all of the unpleasant sounds with any settings of C_2 and C_7, and you have your r.f. stage neutralized. Connect the antenna, and peak C_2 and C_7 on the first signal you find. Do all of your tuning with your regular receiver, and only use C_2 and C_7 to peak the signal when you make a big frequency excursion. The adjustable antenna coupling provides some measure of gain control for the unit, but it is generally best to use fairly tight coupling and hold the gain down in your regular receiver. The antenna coupling is designed for low-impedance input, and will work satisfactorily with

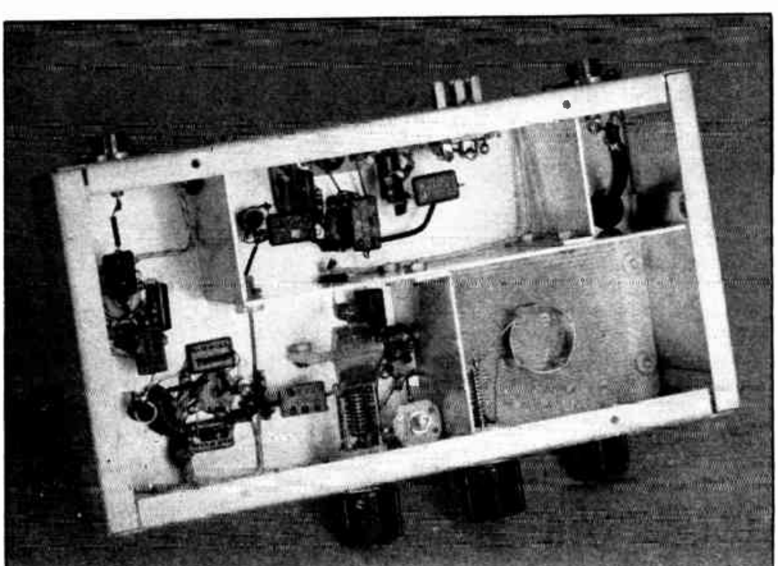


Fig. 5-65 — This view of the underside of the converter with the bottom cover removed shows the Faraday shield at the lower right, the shield straddling the r.f. amplifier socket (lower center) and the shielded oscillator section (top center). The neutralizing condenser for the r.f. stage is adjusted through a hole in the bottom cover.

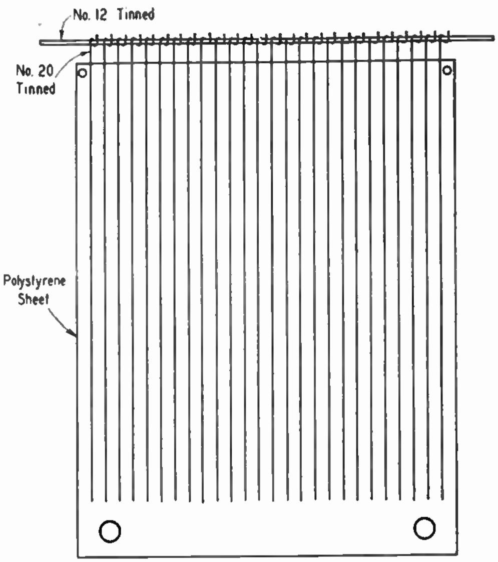


Fig. 5-66 — Constructional details of the Faraday shield, before soldering the ends of the No. 20 wires to the No. 12 wire bus.

50- or 75-ohm line. If you use 300-ohm Twin-Lead, it is better to leave the short length of coaxial line ungrounded and to use something other than a coaxial fitting for connecting the antenna. If your antenna uses 600-ohm line or tuned feeders, it is best to use a small antenna tuning unit link-coupled through a length of RG-59/U to the converter input.

There is nothing sacred about the crystal frequencies used, other than to be sure that they have no harmonics falling within the signal-frequency range. For the crystals suggested in the coil table, the receiver tunes from 4 to 3.6 to cover 14 to 14.4 Mc. (yes, it tunes backwards!), 3.375 to 3.825 for 21 to 21.45 Mc., and 3.5 to 5.2 for 28 to 29.7 Mc. The 27-Mc. amateur band is also covered by the 10-meter converter, simply by tuning your receiver below 3.5 Mc.

What first i.f. (tuning range of your receiver) you will use depends on the available crystals and the range your present receiver tunes. Using the second or third harmonic of the crystal should be satisfactory in practically every case. By careful selection of crystal frequencies, you can arrange things so that the band edges start at some even 100-kc. mark on your receiver, thus giving you frequency-calibrated reception (with the necessary mental correction factor). The accuracy of calibration of your receiver on the one tuning range, together with the accuracy of the crystal used in the oscillator portion of the converter, will determine the accuracy of calibration of the receiving system.

Power Supply

The circuit diagram of a suitable power supply for use with the converters is shown in Fig. 5-67, although any source of 6.3 volts a.c. and 105 and 180 volts d.c. will do. One set of connections runs to the converter in use, and the other goes to a small control box located on the operating table. If desired, the a.c. switch can be incorporated in the power supply, but the plate switch, in the 105-volt lead to the r.f. stage, should be handy to the operator. A switch can be provided for shifting the power from one converter to another. Since separate receiving antennas are generally used at these frequencies, the antennas do not require switching.

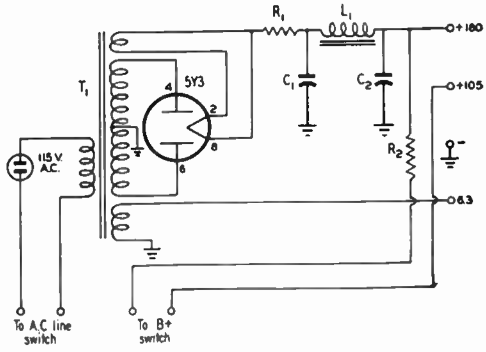


Fig. 5-67 — A power supply for the crystal-controlled converter.
C_1, C_2 — 8-μfd. 450-volt electrolytic.
R_1 — 1500 ohms, 10 watts.
R_2 — 10,000 ohms, 10 watts.
L_1 — 16-hy. 50-ma. choke (Stancor C-1003).
T_1 — 240-0-240 at 40 ma., 5 and 6.3 v. (Stancor P-6297).

A Simple Narrow-Band FM Adapter

Quite a few amateurs are now using NFM transmission, but most of the receivers in current use are of the straight AM type. Reception on a receiver equipped with an FM adapter is quite an improvement, from the standpoint of readability, over the same signal received by detuning the AM receiver to detect the FM signal on the i.f. slope.

With the adapter the "on-signal" noise level from external noise is reduced because of limiter action, and an improvement in readability is immediately noticed when receiving FM signals. Since the adapter allows you to tune to the center of the incoming carrier, any a.v.c. action in the receiver can be used to advantage to hold the "on-signal" noise level down by reducing receiver gain, an advantage that cannot be realized with i.f. slope detection. "Off-signal" noise is somewhat greater than with AM, but this is not too serious since most tuning is done on AM and the adapter is switched in when an FM signal is present.

Basically, the adapter unit consists of a limiter stage followed by a discriminator. The limiter uses a 6SJ7 tube with a 10-$\mu\mu$fd. coupling condenser and a 1-megohm grid leak, as shown in Fig. 5-68. The tube will reach full limiting at about 2.5 microvolts input to the average receiver, and the limiter output is constant over a wide range. Generally, the receiver i.f. system overloads before it is possible to overload the limiter, indicating that little would be gained from the standpoint of maintaining constant output by adding another limiter stage.

The discriminator transformer has two separate low-impedance primary windings, each with a separate secondary winding coupled to it. Each half of the transformer secondary is fixed-tuned with a 510-$\mu\mu$fd. silvered-mica condenser, and variable tuning is accomplished by means of movable iron cores. One of the transformer secondaries is tuned to a frequency approximately 10 kc. higher than the i.f. of receiver to which it is attached. The other is tuned approximately 10 kc. lower than the i.f. The transformer should be used with a receiver having an i.f. of approximately 456 kc. The bandwidth of the transformer is approximately 20 kc. and the characteristic is quite linear over approximately 12 kc. The output of the discriminator is fed directly to the receiver audio system.

Construction of the unit is relatively simple, as will be apparent from reference to Figs. 5-69 and 5-70. It is possible to construct it in an evening. The chassis, which measures 2¼ × 5 × 1 inches, is constructed from a piece of 0.062-inch aluminum sheet measuring 4¼ × 7 inches. Construction can be simplified by

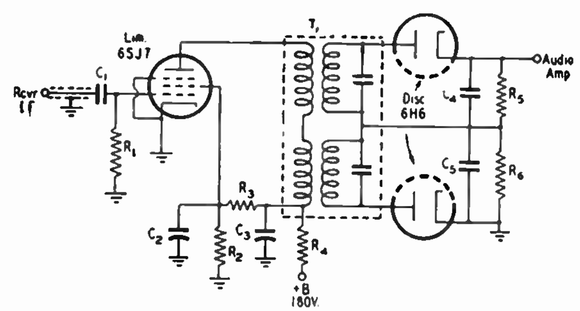


Fig. 5-68 — The NFM-adapter circuit.
C_1 — 10-$\mu\mu$fd. ceramic or mica.
C_2, C_3 — 0.1-μfd. paper.
C_4, C_5 — 100-$\mu\mu$fd. ceramic or mica.
R_1 — 1 megohm, ½ watt.
R_2, R_3 — 33,000 ohms, 1 watt.
R_4, R_5, R_6 — 0.1 megohm, ½ watt.
T_1 — Discriminator transformer (National SA-4842).

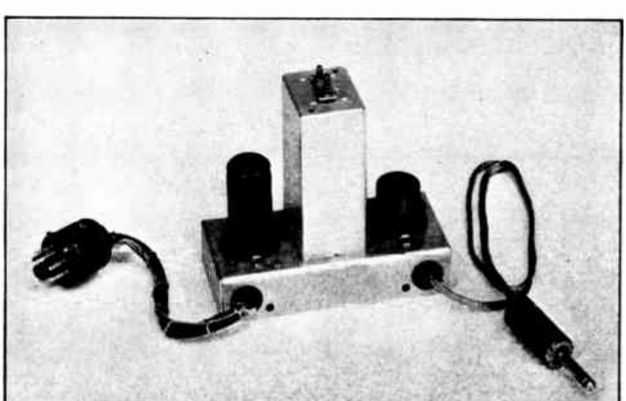


Fig. 5-69 — This two-tube adapter gives NFM reception with a communications receiver having an i.f. in the vicinity of 456 kc. The tubes are a 6SJ7 and a 6H6. The 'phone plug, connected to the audio output terminal in the unit, is plugged into the 'phone jack on the receiver for FM reception, and simply pulled out of the jack for AM.

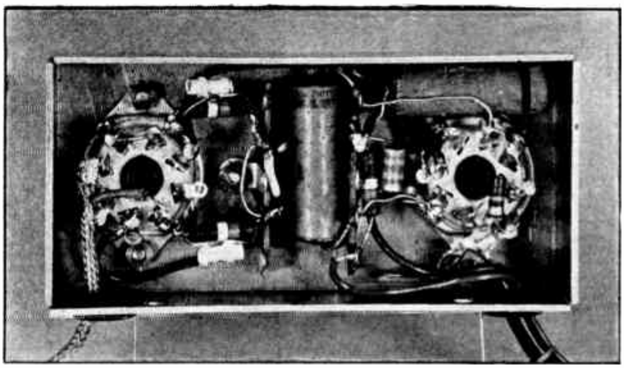


Fig. 5-70 — The simplicity of the wiring is evident in this underneath view of the adapter. The i.f. lead, a piece of small coaxial cable, is laced with the power and audio leads.

using a piece of metal 4¼ by 5 inches and by putting only two bends in the chassis, making it "U"-shaped. Socket holes, as well as the mounting holes for the discriminator transformer, are punched before the chassis is bent to its final shape. Insulated lugs can be mounted on socket screws, as shown, to provide for neat layout of parts.

Some receivers are provided with adapter sockets at the rear into which the adapter may be plugged. This adapter socket provides all voltages necessary to operate the adapter. The audio output can be run through a shielded lead to the phonograph-input jack on the front of the receiver, if the receiver has one, and switching from AM to FM reception is then accomplished by inserting the plug in the phono jack. Simpler methods can be devised, especially if the user has no objection to adding a switch to the front panel of his receiver. In this event it is simply necessary to switch either the AM-detector output or the FM-discriminator output to the audio input of the receiver. It is not necessary to switch off the B-plus of the adapter tubes since interference from cross-talk is negligible.

The i.f. output can be taken from the plate of the second i.f. tube, and there will be some detuning of the detector input transformer. The simplest way to retune the detector transformer, when no signal generator is available, is to set the receiver for maximum background noise with no signal present. When this unit is used in connection with receivers having high-impedance i.f. systems, care must be taken to have the lead from the i.f. tube to the limiter well shielded and as short as possible. Small coaxial cable, such as RG-59/U, should be used for this lead. In the event oscillation troubles are encountered in the receiver with the adapter in place, make sure the adapter unit is well grounded to the receiver and all shielding is attached to a good ground, preferably in the receiver if possible.

High-Frequency Transmitters

Transmitters for the amateur bands lying between 1.8 and 30 Mc. may take a variety of forms, depending primarily upon the frequency bands to be covered and the power output desired. Added to these are such important factors as operating convenience and space restrictions.

The principal requirement that must be met in c.w. transmitters, to which this chapter is limited, is that the output must be confined as closely as the state of the art permits to a single steady frequency free from modulation. A frequency-stable signal is necessary not only to comply with FCC regulations, but also to provide a signal that can be received satisfactorily with a selective receiver, and one that will cause a minimum of interference to amateurs working in the same band. Radiation of signals at harmonic frequencies, or spurious radiations at other frequencies, must be minimized to prevent interference to other radio services, especially television.

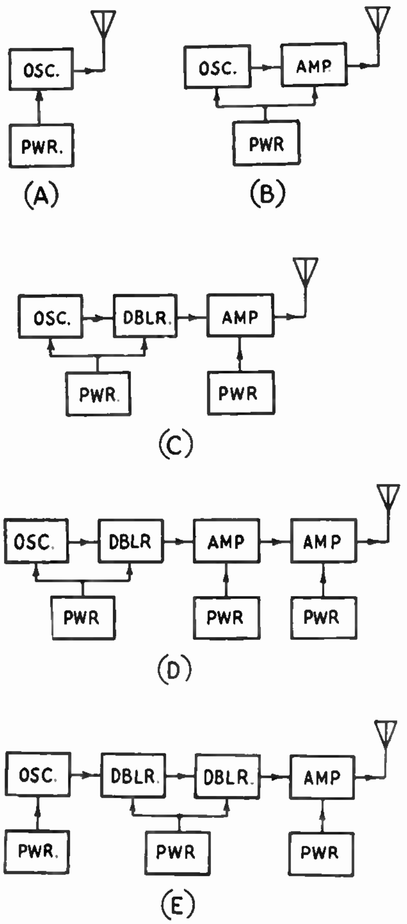


Fig. 6-1 — Block diagrams showing typical combinations of oscillator and amplifiers and power-supply arrangements for transmitters. A wide selection is possible, depending upon the number of bands in which operation is desired and the power output.

A simple oscillator may be used as a transmitter, as shown in Fig. 6-1A, but the amount of power obtainable with satisfactory frequency stability is small. Therefore in most transmitters the oscillator is used to feed one or more amplifiers as required to bring the power up to the desired level, as indicated at B, before delivering the power to the antenna system.

An amplifier whose output frequency is the same as the input frequency is called a **straight amplifier**. If such a straight amplifier is placed in an intermediate position between two other transmitter stages it is sometimes called a **buffer amplifier.**

Because it becomes increasingly difficult to maintain oscillator frequency stability as the frequency is increased, it is most usual practice in working at the higher frequencies to operate the oscillator at a low frequency and follow it with one or more **frequency multipliers** as required to arrive at the desired output frequency. A frequency multiplier is an amplifier that delivers output at a multiple of the exciting frequency. A **doubler** is a multiplier that gives output at twice the exciting frequency; a **tripler** multiplies the exciting frequency by three, etc. Although multiplications in a single stage as high as eight or more sometimes are used to reach the bands above 30 Mc., in the majority of low-frequency transmitters, multiplication in a single stage is limited to two or three, since the efficiency of a multiplier decreases rapidly as the order of multiplication increases. Also, it becomes more difficult to keep unwanted harmonics from the output.

Frequency multipliers sometimes are used to feed the antenna system directly, but preferably should feed a straight amplifier which, in turn, feeds the antenna system, as shown in Fig. 1-C, D and E, because it is otherwise difficult to eliminate the multiplier driving frequency and undesired harmonics in the antenna system. As the diagrams indicate, it is often possible to operate more than one stage from a single power supply.

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 megacycle) 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 (or multiples of that frequency). By making the inductance, capacitance, or both, variable in the self-controlled oscillator, it may be operated at any frequency desired within a band 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 is preferred by many others because of the comparative ease with which frequency stability and calibration are maintained.

While any of the basic self-controlled oscillator circuits may be used, the prevailing choice lies among those shown in Fig. 6-2, or modifications of these circuits.

To provide satisfactory performance on the air, special attention must be paid to the circuit and mounting of parts. 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 perhaps 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.

Effects of Load

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 capacitances. For this

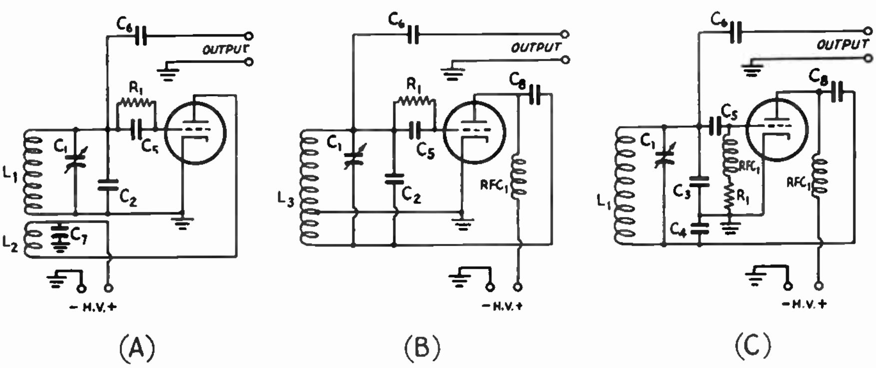


Fig. 6-2 — Typical simple VFO circuits. A — Tickler feed-back. B — Hartley. C — Colpitts. Low-μ triodes, such as the 6C5 or 6J5, are preferable. Approximately appropriate values for the 3.5-Mc. band are as follows:

C_1 — Tuning condenser — 150-μμfd. variable.
C_2 — Tank condenser — 500-μμfd. zero-temp. mica.
C_3 — Tank condenser — 700-μμfd. zero-temp. mica.
C_4 — Tank condenser — 0.0021-μfd. zero-temp. mica.
C_5 — Grid condenser — 100-μμfd. zero-temp. mica.
C_6 — Output coupling condenser — 100 μμfd. or less, mica.
C_7 — Plate by-pass condenser — 0.01-μfd. paper.
C_8 — Plate blocking condenser — 0.001-μfd. mica.
R_1 — Grid leak — 50,000 ohms.
L_1 — Tank coil — 4.3μh.
L_2 — Tickler winding — Approximately one-third number of turns on L_1, wound on same form next to L_1 or over ground end of L_1.
L_3 — Same as L_1, tapped approximately one-third from plate end.
RFC_1 — Parallel-feed r.f. choke — 2.5 mh.

reason it is almost universal practice to use an amplifier between the VFO and the antenna system.

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. Special oscillator circuits have been developed to minimize this effect. Two forms of the **electron-coupled** oscillator circuit are shown in Fig. 6-3. In circuits of this type a single screen-grid tube performs the functions of both an oscillator and an amplifier. The screen serves as the plate of a triode oscillator, while the power is taken from a separate tuned output-plate tank circuit, the coupling between the two being principally through the common electron stream.

In Fig. 6-3A, the oscillator circuit is a Hartley in which the ground point has been shifted from the cathode to the "plate." Fig. 6-3B shows the Colpitts modified in a similar manner. The choke, RFC_1, is required to provide a d.c. path to the cathode without grounding it for r.f.

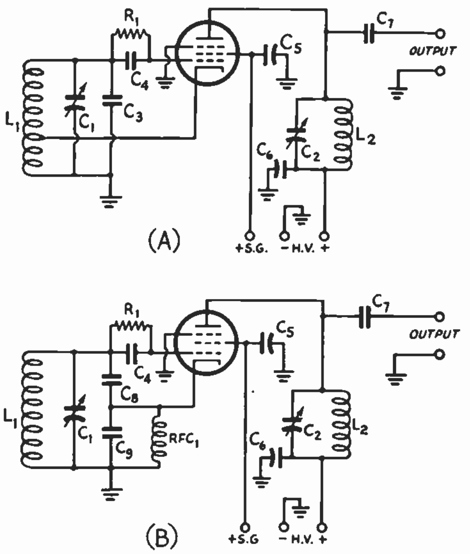


Fig. 6-3 — ECO circuits. A — Hartley. B — Colpitts. Approximate values are as follows:

C_1 — Oscillator tuning condenser — for 3.5 Mc.: 150-μμfd. variable.
C_2 — Output tank condenser — 100-μμfd. variable.
C_3 — Tank condenser — 500-μμfd. zero-temp. mica for 3.5 Mc.
C_4 — Grid condenser — 100 μμfd. or less, zero-temp. mica.
C_5 — Screen by-pass condenser — 0.01-μfd. paper.
C_6 — Plate by-pass condenser — 0.01-μfd. paper.
C_7 — Output coupling condenser — 100 μμfd. or less, mica.
C_8 — 700-μμfd. zero-temp. mica.
C_9 — 0.0021-μfd. zero-temp. mica.
R_1 — Grid leak — 50,000 ohms.
L_1 — Oscillator tank coil — 4.3 μh. tapped approximately one-third from ground end for 3.5 Mc.
L_2 — Output tank coil — 22 μh. for 3.5 Mc., 7.5 μh. for 7 Mc.
RFC_1 — Parallel-feed r.f. choke — 2.5 mh.

In both of these circuits, output at a multiple of the oscillator frequency may be obtained by tuning the output-plate tank circuit to the desired harmonic, although this is seldom done beyond the second harmonic.

The oscillator frequency is not entirely independent of tuning or loading in the output plate circuit. The reaction is less, however, when the output-plate circuit is tuned to a harmonic or replaced by an untuned circuit, such as an r.f. choke, as shown in Fig. 6-4. The power output obtainable with the latter arrangement is much lower, however.

Well-screened tubes are preferable as electron-coupled oscillators. Those commonly used are the 6K7, 6SK7, 6F6 or 6AG7.

Another measure that may be taken to provide isolation between the oscillator and a following tuned amplifier is the use of an untuned amplifier, as shown in Fig. 6-5.

The power gain of an amplifier of this type is quite small, the purpose being almost entirely that of securing isolation between the VFO and tuned power amplifiers whose adjustment might react on the frequency of the oscillator if coupled to it directly. Two amplifier stages of this type usually are necessary before a following amplifier can be tuned or keyed without noticeably affecting the oscillator frequency and stability.

When using such an amplifier following an electron-coupled oscillator, a nonresonant output circuit also is usually used in the ECO. R.f. chokes are used as nonresonant circuits in the outputs of the ECO and in the second amplifier. L_1 in the plate circuit of the first amplifier is a winding that is self-resonant with the tube and circuit capacitances at a frequency near but not in the band of frequencies over which the amplifier is intended to operate. This is to prevent forming a low-frequency t.g.t.p. oscillating circuit which occurs when chokes of approximately the same characteristics are used in both input and output circuits of the amplifier tubes. For the same reason, resistors without chokes are used in the grid circuits.

Regulated voltages for the screens and plates are desirable.

Chirp

Variations in the voltage of the oscillator-tube elements can 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 regulated plate and screen voltages. 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, when the key is open, and full voltage, when the key is closed, 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, unless the oscillator can be made insensitive to voltage changes by other means. 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 change 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 of good construction and 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.

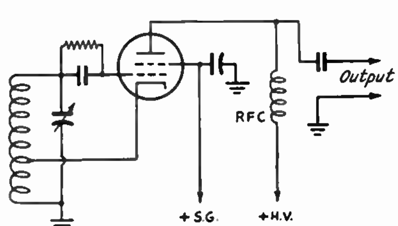


Fig. 6-4 — ECO with an r.f. choke replacing the output tank circuit for the purpose of reducing reaction on the oscillator portion of the circuit.

Mechanical Considerations

Any mechanical vibration which causes a change in the capacitance across the tank circuit, or in dimensions of the coil, will cause a corresponding change in frequency. This should be minimized by solid construction, by 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.

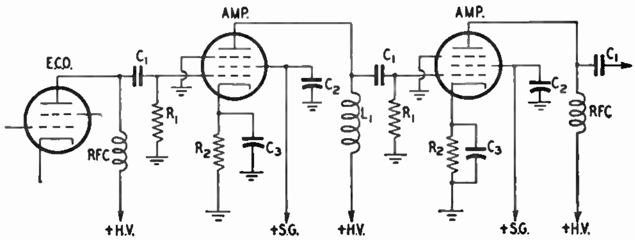


Fig. 6-5 — Diagram showing two isolating-amplifier stages coupled to the output of an ECO. Well-screened tubes are preferable, 6K7s or 6F6s being suitable.

C_1 — Coupling condenser — 100 $\mu\mu$fd. or less, mica.
C_2 — Screen by-pass condenser — 0.01-μfd. paper.
C_3 — Cathode by-pass condenser — 0.01-μfd. paper.
R_1 — Grid leak — 50,000 to 100,000 ohms.
R_2 — Cathode biasing resistor — 200 to 500 ohms.
L_1 — Coupling inductance — see text.
RFC — Plate choke — 2.5 mh.

Plug-in coils for changing oscillator frequency ranges are not recommended because experience has shown that the coil contacts may become the source of undesirable frequency instability.

VFO Tank Q

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 capacity changes a small percentage of the total circuit capacitance. 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. An increase in Q can be obtained also by tapping the tube across only a portion of the tank circuit. Fig. 6-6A shows the Hartley circuit with the grid and plate tapped across small portions of the tank-coil reactance. An equivalent arrangement for the Colpitts circuit is shown at B. C_1 (and C_2 in parallel) is small compared with C_3 and C_4. Therefore, the reactance across which each tube element is connected is a small portion of the total. C_2, which is the tuning condenser, should be no larger than is necessary to tune across the band so as not to influence the function of C_1 any more than necessary. The tuning condenser should not be connected across the coil, since this reduces the Q of the circuit.

In both of these arrangements, the higher the Q of the coil, the smaller the reactance between tube elements may be without stopping oscillation and, therefore, the greater the stability. Because of the high L/C ratio which results with the circuit of B, greater care must be exercised in the construction and mounting of tank-circuit components.

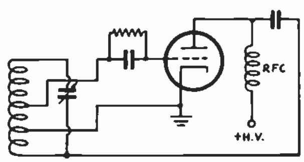
Any of the bandspread tuning systems used in receivers may be applied to the oscillator circuits which have been under discussion. The parallel-condenser system is used most widely since it lends itself well, particularly to high-C circuits.

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.

VFO ADJUSTMENT

Tuning Characteristics

Normally-operating VFO circuits of the types under discussion 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 a grounded prod 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 and screen voltages, 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.



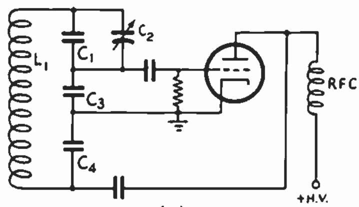


Fig. 6-6 — As an alternative to the use of a high-C tank circuit, oscillator tubes sometimes are connected across only a portion of the tank circuit to increase the Q. In the Hartley circuit of A, the grid and plate connections are made to taps instead of to the ends of the coil. In the Colpitts circuit of B, the division is by capacitive means. For 3.5 Mc., C_3 and C_4 should be about 0.001 μfd. and $C_1 + C_2$ no larger than necessary to maintain oscillation and tune across the band. The Q of L_1 and the G_m of the tube should be as high as possible.

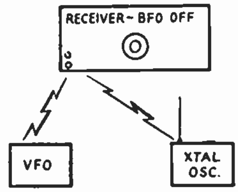


Fig. 6-7 — Set-up for checking VFO stability. The receiver should be tuned preferably to a harmonic of the VFO frequency. The crystal oscillator may operate somewhere in the band in which the VFO is operating. The receiver b.f.o. should be turned off.

In the circuit of Fig. 6-6A, maximum frequency stability is obtained with the plate and grid taps as close as possible to the cathode tap without stopping oscillation. In the circuit of Fig. 6-6B, maximum stability is obtained when C_3 and C_4 (usually equal) are large and the ratio of $C_1 + C_2$ to C_3 or C_4 is the maximum possible without stopping oscillation. The adjustment in each case will be limited by the Q of the coil. Therefore, the Q must be high for greatest frequency stability.

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, well-shielded monitoring arrangements is a receiver combined with a crystal oscillator, as shown in Fig. 6-7. (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 at the lower frequencies will be magnified at the higher frequencies, accurate checking can best be done by monitoring at the latter.

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

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. The oscillator normally should be considered as a frequency-generating device only, with power output of secondary importance. The amount of power which may be obtained from a crystal oscillator is limited by the heat the crystal will stand without fracturing. The amount of heating is dependent upon the r.f. crystal current which, in turn, is a function of the amount of feed-back required to provide proper excitation. Crystal heating short of the danger point results in frequency drift to an extent depending upon the way the crystal is cut. Excitation should always be adjusted to the minimum necessary for proper operation.

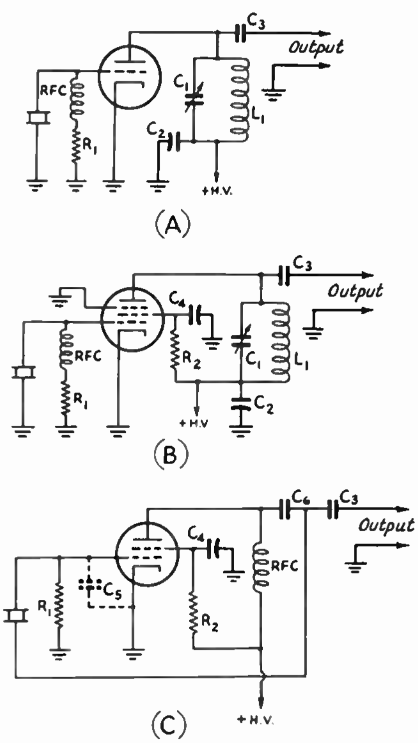


Fig. 6-8 — Simple crystal-oscillator circuits. A — Triode. B — Tetrode or pentode. C — Tetrode Pierce. Approximate values are as follows:

C_1 — Tank condenser — 100-$\mu\mu$fd. variable.
C_2 — Plate by-pass condenser — 0.01-μfd. paper.
C_3 — Output coupling condenser — 100 $\mu\mu$fd. or less, mica.
C_4 — Screen by-pass condenser — 0.01-μfd. paper.
C_5 — Feed-back condenser — 50 to 100 $\mu\mu$fd.
C_6 — Plate blocking condenser — 0.001-μfd. 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 μhy. for 3.5 Mc.; 7.5 μhy. for 7 Mc.
RFC — Parallel-feed r.f. choke — 2.5 mh.

SIMPLE CIRCUITS

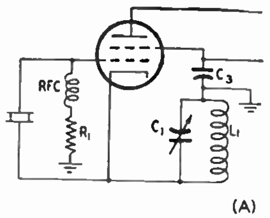
The basic crystal-controlled oscillator circuits are shown in Fig. 6-8. Since the crystal is the equivalent of a high-Q tuned circuit of fixed frequency, it will be observed that each of the crystal circuits is essentially the equivalent of a self-controlled circuit.

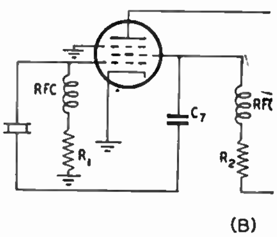
Triode, Tetrode and Pentode Oscillators

The triode crystal circuit of Fig. 6-8A is the equivalent of the t.g.t.p. circuit in which a 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 being grounded as shown. The same circuit is used for tetrodes, such as the 6V6 and 6L6, the suppressor connection being omitted.

With this circuit, oscillation takes place only when the plate tank circuit is tuned to a frequency higher than that of the crystal, and maximum output usually occurs when it is tuned close (but not exactly) to the crystal 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 jump to a relatively high value, as shown at the left in Fig. 6-9. As the plate circuit is tuned past the point of resonance with the crystal 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

HIGH-FREQU





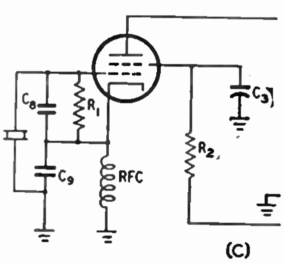


Fig. 6-10 — Crystal-controlled circuits. A — Tri-tet. B and C — proximate values are as follows:
C_1 — Cathode-tank tuning conder
C_2 — Output-tank tuning condens
C_3 — Screen by-pass condenser — 0.01-μfd. paper.
C_4 — Plate by-pass condenser — (
C_5 — Output coupling condenser -
C_6 — Feed-back control condenser
C_7 — Parallel-feed blocking conde
C_8 — Feed-back-adjustment conde
C_9 — Feed-back-adjustment cond $\mu\mu$fd.
R_1 — Grid leak — 50,000 to 150,0(
R_2 — Screen voltage-dropping r 100,000 ohms.
L_1C_1 — Tuned well above crystal f
L_2C_2 — Tuned to crystal frequency
RFC — Parallel-feed r.f. choke — :

When operating the Tri-t crystal frequency, L_1C_1 shoi closer to the crystal freque essary to make the circuit output-plate voltage appliec screened tubes, it may be ad circuit L_1C_1 when operating fundamental frequency, revert circuit as a measure of safety

With well-screened tubes, st 2E25 or 802, the output-plat teristic is like that shown in fundamental as well as at the

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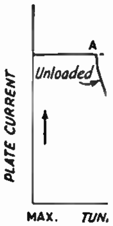

Fig. 6-9 — General tu
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resonance and that a weak self-oscillation will occur instead. This condition can usually be avoided by proper proportioning of the two feed-back capacitances. If fixed capacitances are to be used, the values given under Fig. 6-10 are suggested, but they may have to be modified somewhat, depending upon the type of tube used.

CRYSTALS

Crystal Characteristics

While crystals are produced for frequencies as high as 50 Mc., by far the majority of those used in amateur high-frequency transmitters are cut for the 3.5- and 7-Mc. bands. With suitable frequency-multiplying stages, this permits the use of a single crystal for operation in the harmonically-related parts of higher-frequency bands, as well as at the crystal fundamental frequency. As an example, a 3501-kc. crystal with appropriate multipliers may be used for frequencies of 7002 kc., 14,004 kc., 28,008 kc., etc.

Crystals vary in characteristics depending upon the manner in which they are cut from the natural quartz crystal, particularly in the thickness-frequency and temperature-frequency relationships. The frequency of crystals of the earlier cuts, designated "X" and "Y," vary appreciably in frequency with changes in temperature. More recently they have been almost entirely superseded by the modern "AT-" and "BT"-type crystals which are cut so as to have very small frequency change with temperature. The temperature-frequency characteristics for various crystal types are summarized in the following tabulation:

X-cut — −10 to −25 cycles per Mc. per degree C.
Y-cut — +100 to −20 cycles per Mc. per degree C.
AT-cut — +10 cycles per Mc. per degree at 0 degrees C.
— 0 cycles per Mc. per degree at 45 degrees C.
— +20 cycles per Mc. per degree at 85 degrees C.
BT-cut — −10 cycles per Mc. per degree at 0 degrees C.
— 0 cycles per Mc. per degree at 30 degrees C.
— −20 cycles per Mc. per degree at 70 degrees C.

The relationship between the thickness of a crystal and its frequency is given by:

$$f_{Mc.} = \frac{k}{t_{mil}}$$

where $f_{Mc.}$ is the frequency in megacycles, t the thickness in thousandths of an inch and k is a constant of the crystal cut approximately as follows:

X-cut — 112.6
Y-cut — 77
AT-cut — 66.2
BT-cut — 100.78

An AT crystal usually is more active than one of the BT-cut type, but since it is thinner for the same frequency, there is greater danger of fracture in operation. Therefore, AT-cut crystals usually are used for frequencies below 5 Mc., while the BT-cut is used for crystals whose frequencies lie above 5 Mc., although this is not true in all cases.

While crystals are sometimes cut for fundamental frequencies as high as 14 Mc., most crystals used by amateurs for frequencies higher than the 7-Mc. band are "harmonic-type" crystals; that is, the thickness corresponds to a frequency of one-third (sometimes one-fifth) of the normal operating frequency. The other dimensions of the crystal are proportioned so that the mechanical vibration is at three times (or five times) the fundamental frequency.

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. 6-12), 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 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. 6-13 is desirable. It consists of an FT243 holder with a sliding cover

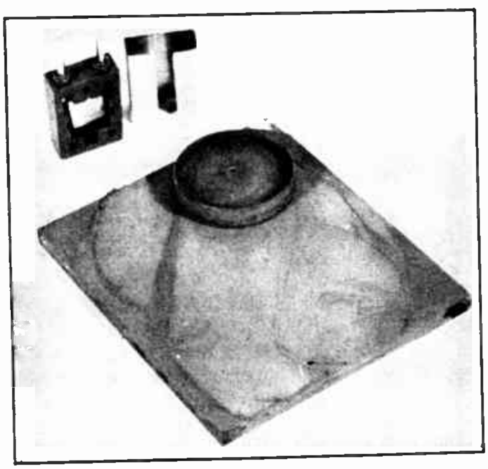


Fig. 6-12 — The equipment necessary for grinding a crystal blank to frequency. A piece of plate glass and a "button" of the same material are essential. The "quick-change" adaptation for the crystal holder is a convenience. Not shown, but also convenient, are a small paint brush for spreading abrasive and a toothbrush for scrubbing.

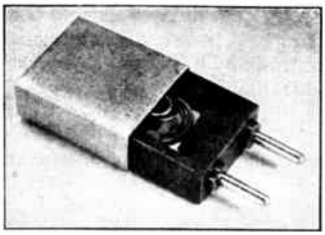


Fig. 6-13 — The quick-change crystal holder with sliding cover.

fashioned from sheet metal. Soap, warm water and a toothbrush are used to clean and rinse 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. 6-14, 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.

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 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.

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.

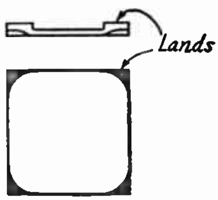


Fig. 6-14 — The ½ × ½-inch electrodes used in modern crystal holders, showing the lands at the corners between which the crystal is firmly held.

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 Amplifier Circuits

The power output from an oscillator is limited for reasons previously stated. Power greater than a few watts 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. 6-15 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 for the tuned-grid tuned-plate oscillator. Therefore the amplifier circuit itself will function as an oscillator, independent of the oscillator feeding it, unless measures are taken to reduce the plate-grid capacitance or nullify its effect.

TRIODE CIRCUITS

Plate Capacitive Neutralizing Systems

The plate-grid capacitance can be **neutralized** by feeding back to the grid, through an external path, a voltage which at any instant is equal, but in opposite phase, to the voltage fed back through the tube.

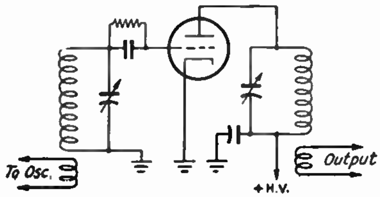


Fig. 6-15 — Fundamental r.f. power-amplifier circuit. Means must be provided to prevent oscillation since the circuit is the same as that for a t.g.t.p. oscillator. See text for discussion.

The most commonly-used circuits for this purpose are shown in Fig. 6-16. 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. Thus the voltages at opposite ends of the tank are essentially equal, but 180 degrees out of phase.

The neutralizing and feed-back voltages are matched in amplitude by adjusting the capacitance of the neutralizing condenser, C_6.

In Fig. 6-16A, 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, including socket and other external stray capacitances across the elements.

In Fig. 6-16B, 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.

Push-Pull Triode Circuits

Fig. 6-16C and D show equivalent push-pull arrangements. In circuit D, better circuit balance can be maintained by using split-stator tank condensers. The rotors in this case should *not* be grounded.

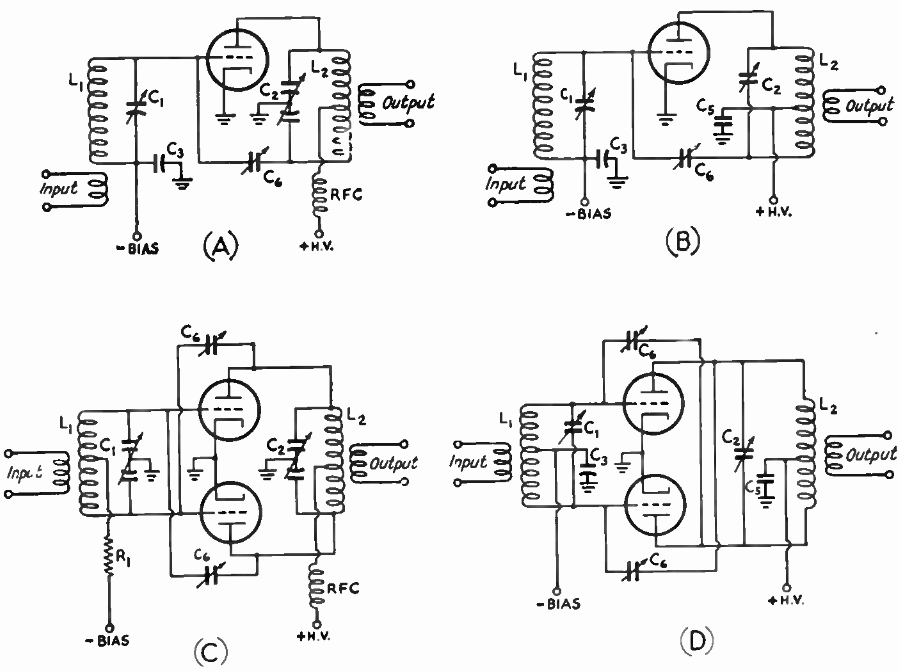


Fig. 6-16 — Neutralized-triode amplifier circuits. A — Single tube with capacitive balance. B — Single tube with inductive balance. C and D show corresponding push-pull arrangements.

C_1L_1 (grid tank) and C_2L_2 (plate tank) are tuned to the frequency fed to the amplifier.

C_3 — Grid by-pass condenser — 0.001-μfd. mica to 0.01-μfd. paper.

C_5 — Plate by-pass condenser, 0.001-μfd. mica to 0.01-μfd. paper.

C_6 — Neutralizing condenser — approximately same capacitance as tube grid-plate capacitance.

R_1 — Grid-circuit isolating resistor — 100 ohms.

RFC — Plate-circuit isolating radio-frequency choke — 1 to 2.5 mh.

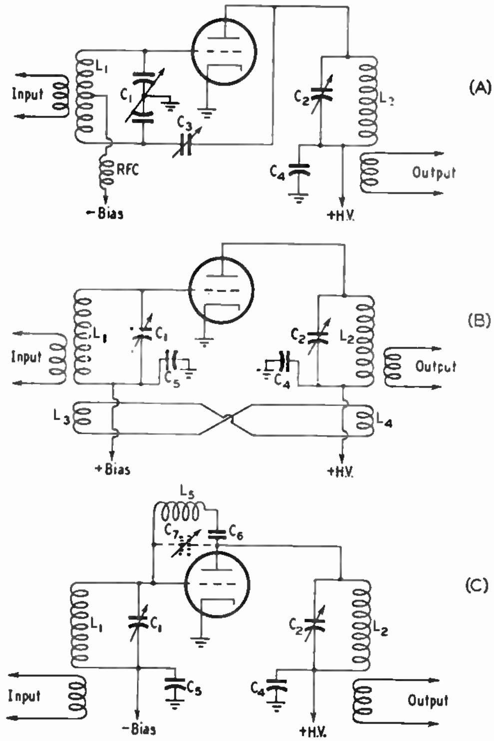


Fig. 6-17 — Additional, but less commonly-used neutralizing circuits. A — Grid neutralizing. B — Link neutralizing. C — Inductive neutralization.

C_1L_1, C_2L_2 — Tank circuits tuned to operating frequency.
C_3 — Neutralizing condenser — approximately same capacitance as grid-plate capacitance of tube.
C_4 — Plate by-pass condenser — 0.001-μfd. mica to 0.01-μfd. paper.
C_5 — Grid by-pass condenser — 0.001-μfd. mica to 0.01-μfd. paper.
C_6 — Voltage-blocking condenser — 0.001-μfd. mica.
C_7 — Variable condenser to tune trap circuit to operating frequency with L_5 and grid-plate capacitance of tube. (See text.)
L_3, L_4 — Neutralizing links — 2 to 10 turns, depending upon frequency.
L_5 — Neutralizing trap coil — to tune to operating frequency with C_7 and grid-plate capacitance of tube. (See text.)

In Fig. 6-16C, the r.f. choke in the plate circuit prevents r.f. grounding of the coil center-tap (through the power supply) and the rotor of the condenser simultaneously. This condition is to be avoided because it sets up three tuned circuits — each half of the tank circuit in addition to the circuit as a whole. The isolating resistor in the grid circuit serves a similar purpose.

Grid Capacitive Neutralizing Systems

Additional, but less widely-used neutralizing circuits are shown in Fig. 6-17. The circuit of Fig. 6-17A is similar to that of Fig. 6-16A, 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 comparable to Fig. 6-16B.

Link Neutralizing Circuit

The link neutralizing circuit of Fig. 6-17B 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 the major alteration in either grid or plate 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.

Inductive Neutralization

The inductive neutralizing arrangement of Fig. 6-17C consists merely of making the plate-grid capacitance of the tube part of a circuit tuned to the frequency at which the amplifier is designed to operate. Since such a circuit presents a high impedance to the flow of current at the frequency to which it is tuned (wavetrap), it prevents feed-back. C_7 may be

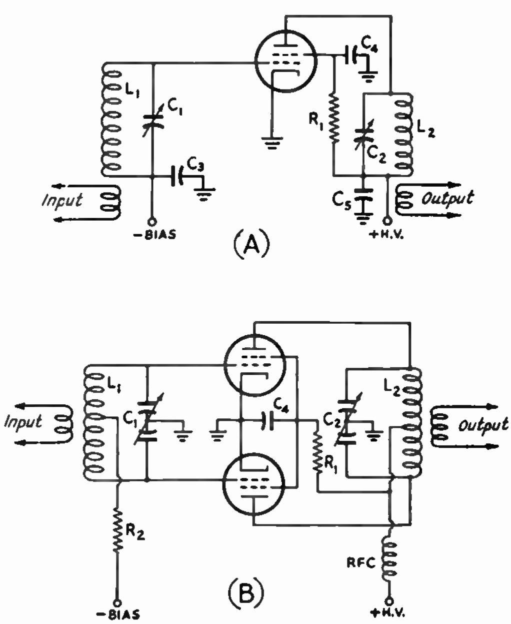


Fig. 6-18 — Screen-grid amplifier circuits. A — Single-tube amplifier. B — Push-pull.

C_4 — Screen by-pass condenser — 0.001-μfd. mica to 0.01-μfd. paper.
R_1 — Screen voltage-dropping resistor.
R_2 — Grid-circuit isolating resistor — 100 ohms.
Other values same as Fig. 6-16.

added for adjustment, although this may decrease the frequency range over which one neutralizing adjustment will hold.

All of the circuits of Fig. 6-17 have disadvantages in amateur practice, particularly in respect to the tuning range over which a single adjustment of neutralization will hold.

SCREEN-GRID AMPLIFIER CIRCUITS

Single-tube and push-pull screen-grid amplifier circuits are shown in Fig. 6-18. The grounded screen in transmitting-type tetrodes and pentodes serves as a shield between the plate and grid to reduce the grid-plate capacitance to the point where feed-back is insufficient to support oscillation. Thus tubes of this type are designed to be operated without neutralization when circuit simplicity is of importance. However, neutralization usually will result in more reliable stability. Poorly-screened audio tetrodes, such as the 6L6 and 6V6, invariably require neutralization.

The power sensitivity of screen-grid tubes is much higher than that of triodes of comparable power rating. Therefore greater care must be exercised in eliminating possible paths for feed-back coupling external to the tube.

Interstage Coupling Systems

Of the various systems that have been devised for feeding the output of one stage into the input of another, the inductive-link and capacitive systems are the most widely used in amateur transmitters. The link system is used principally in cases where there must be appreciable physical separation between stages, where balanced and unbalanced circuits are to be coupled, or when minimum circuit capacitance is desired. Coupling is adjusted more readily with this system and harmonic energy is not so easily coupled from stage to stage. Therefore link coupling has considerable advantage when TVI is a consideration. The capacitive system has the advantages of simplicity, cheapness and compactness, but it does not lend itself so readily to the conditions listed above.

INDUCTIVE COUPLING SYSTEMS

Link Coupling

The link system, examples of which are shown in Fig. 6-19, consists merely of a two-wire low-impedance line with each end terminated in a coil of a few turns coupled tightly to the low-potential point of the output tank coil of the driver and the input tank coil of the driven stage. This low-potential point occurs at the "ground" end of the tank coil in unbalanced circuits (Fig. 6-19A, B and C) and at the center of the tank coil in balanced arrangements (Fig. 6-19B, C and D).

The coupling between the two stages is largely a matter of the tank-circuit Qs but it can be adjusted within limits either by changing the number of turns in the link windings or by changing the coupling between the links and the tank coils. If increasing the number of link turns does not provide sufficient coupling, the tank-circuit Q must be increased.

Fig. 6-19A shows the link system coupling two unbalanced circuits. This arrangement would be used, for instance, in coupling an oscillator or a screen-grid driver to the input of a single-tube stage.

The scheme at B would be suitable for coupling a neutralized or push-pull driver to a single-tube amplifier.

Fig. 6-19C shows the method applied in coupling the output of an unneutralized driver to a push-pull amplifier, while D is the circuit to be used in coupling a neutralized or push-pull stage to another push-pull input.

Inductive Coupling

Another system which is used sometimes in coupling between an unbalanced driver and a balanced amplifier is shown in Fig. 6-20. The output coil of the driver stage is designed to resonate, with the driver-tube and stray circuit capacitances, near the desired operating frequency. The amplifier input tank circuit tunes to the operating frequency and serves to a considerable degree also to tune the output circuit of the driver stage, since the two coils are coupled quite tightly. L_1 should be wound centrally over or inside L_2 and the turns of L_1 adjusted experimentally for optimum power transfer. Sometimes both circuits are tuned, in which case the coils need not be coupled so tightly.

CAPACITIVE COUPLING CIRCUITS

In a capacitive coupling system, the output tank circuit of the driver stage serves also as the input tank circuit of the driven stage. Several arrangements for coupling between balanced and unbalanced circuits, depending upon whether series or parallel power feed is desired, are shown in Fig. 6-21.

With capacitive coupling, the two stages cannot be separated physically any appreciable distance without involving loss in transferred power, radiation and the danger of instability because of feed-back which long high-impedance leads may provide. Since both the output capacitance of the driver tube and the input capacitance of the driven tube are lumped across the single tuned circuit, this sometimes makes it difficult, with the high-capacitance of screen-grid tubes, to obtain a tank circuit with a sufficient amount of inductance to provide an efficient circuit for the higher frequencies. Another disadvantage is that it is difficult to preserve circuit balance in

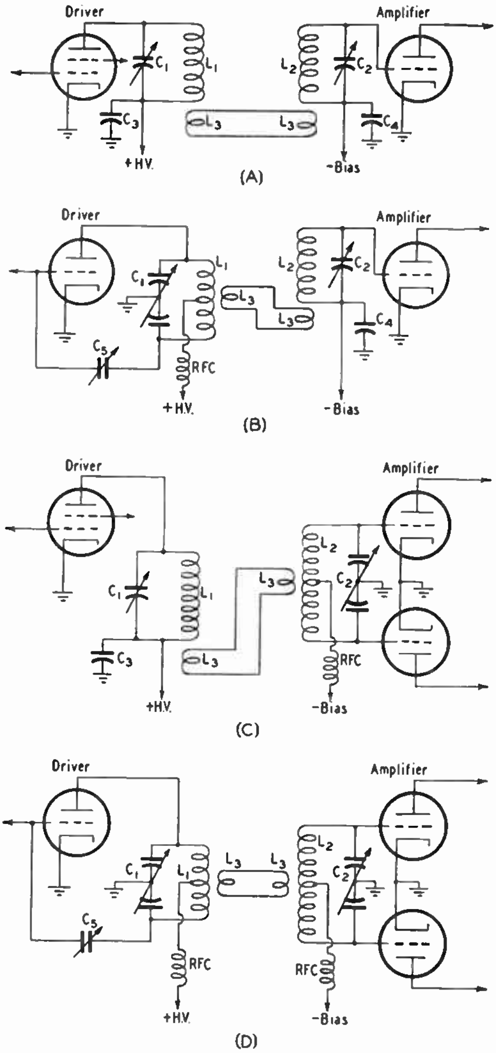


Fig. 6-19 — Link coupling circuits. A — Unbalanced output to unbalanced input. B — Balanced output to unbalanced input. C — Unbalanced output to balanced input. D — Balanced output to balanced input.

C_1 — Driver-stage plate tank condenser.
C_2 — Driven-stage grid tank condenser.
C_3 — Plate by-pass condenser.
C_4 — Grid by-pass condenser.
C_5 — Neutralizing condenser.
L_1 — Driver output tank coil.
L_2 — Driven-stage input tank coil.
L_3 — Link winding.
C_1L_1 and C_2L_2 are always tuned to the same frequency.
RFC — R.f. choke.

coupling from a single-tube stage to a push-pull stage because the circuit tends to become unbalanced by the output capacitance of the driver tube which appears across only one side of the circuit. This does not, however, preclude its use for this purpose, if simplicity in circuit is considered of greater importance, for frequencies below 30 Mc.

The arrangements of Fig. 6-21A and B are most often seen with the plate tap of A and the grid tap of B connected to the top end of the coil. A is used when series driver plate feed is desired; B when series amplifier grid feed is wanted. In the circuit of C, the tank condenser and coil are grounded directly, but parallel power feed is required for the driver plate and usually for the amplifier grid although the grid leak sometimes is placed across the coupling condenser, C_3.

An arrangement which makes possible series feed to both plate and grid is shown at D. L_1 in D is a single coil, opened at the center for feeding in plate and biasing voltages. Since the by-pass condensers, C_2, are directly in the tank circuit, they should be of good-grade mica and capable of handling the r.f. current circulating through the tank circuit. The scheme is practical chiefly in low-power stages. Because it provides a "double-ended" output circuit, it may be used in a neutralized amplifier stage simply by the addition of neutralizing condenser C_5. The grid of the driven tube and the plate of the driver tube being connected across opposite halves of the tank circuit helps to distribute stray capacitances more evenly, thereby preserving a better circuit balance. A still better balance can be achieved by using a split-stator condenser at C_1 and a single mica condenser at C_2, grounding the circuit at the split-stator rotor rather than between the two fixed condensers. Excitation may be adjusted, if necessary, by tapping the grid or plate, as may be required, down on the coil. Such a change, however, will necessitate readjustment of neutralization if the tank is used for neutralizing the driver as suggested.

The circuit of Fig. 6-21E is the preferred arrangement for coupling a neutralized driver to a single-tube amplifier in cases where series feed to the grid of the amplifier is not considered important. F shows the same system feeding a push-pull amplifier. If a more accurate balance is desired, a balancing condenser, C_6, can be used across the other half of the circuit to compensate for the driver-tube output capacitance.

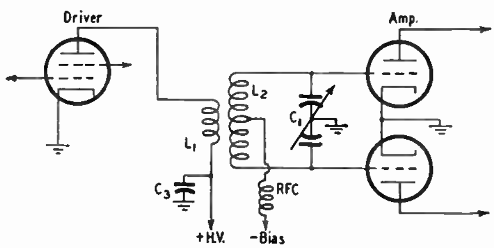


Fig. 6-20 — Inductive coupling from unbalanced output to balanced input.

C_1 — Driven-stage grid tank condenser.
C_3 — Plate by-pass condenser.
L_1 — Self-resonant (approximately) output coil.
L_2 — Driven-stage grid tank coil.
L_1 and L_2 should be coupled tightly.
RFC — R.f. choke.

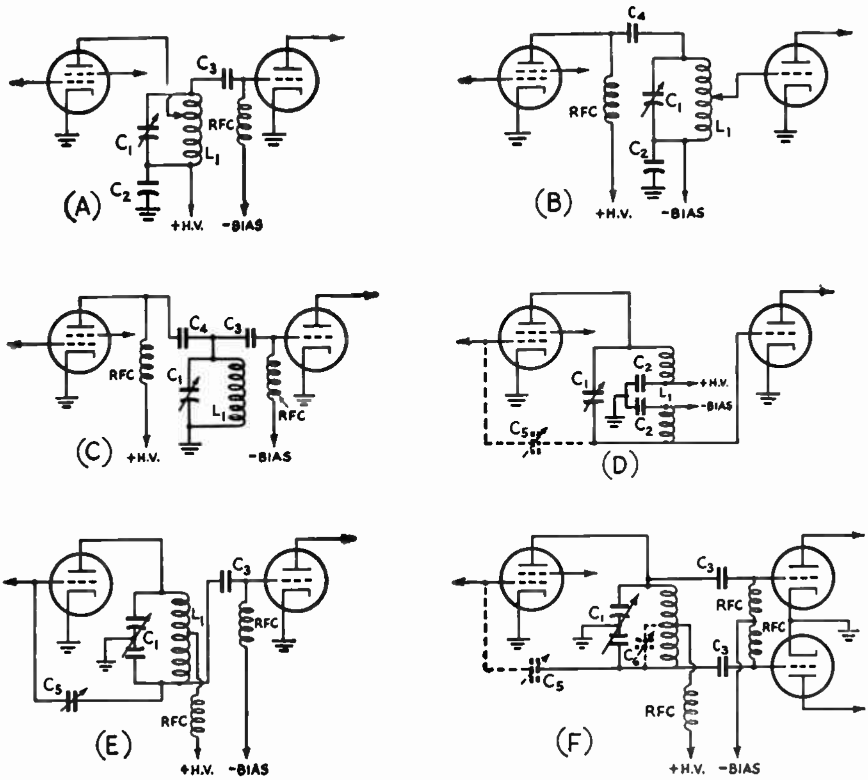


Fig. 6-21 — Examples of capacitive coupling. A — Series plate feed, parallel grid feed. B — Parallel plate feed, series grid feed. C — Parallel feed in both plate and grid. D — Series feed in both plate and grid. E — Balanced output to unbalanced input, series plate feed, parallel grid feed. F — Single tube to push-pull.

C_1 — Tank condenser.
C_2 — By-pass condenser.
C_3 — Coupling condenser.
C_4 — Driver plate blocking condenser.
C_5 — Driver neutralizing condenser.
C_6 — Circuit-balancing condenser.
L_1 — Tank coil.
RFC — R.f. choke.

Capacitive-Coupling Adjustment

Overcoupling can be remedied by reducing the capacitance of the coupling condenser, or by tapping the grid of the driven tube across only a portion of the tank coil, as indicated in Fig. 6-21B. If increasing the capacitance of the coupling condenser does not provide sufficient coupling, the tank-circuit Q must be increased. This can be done by decreasing the L/C ratio of the tank circuit or by tapping the plate of the driver across a portion of the tank coil, as shown in Fig. 6-21A. However, it is preferable and often possible to choose a tank-circuit L/C ratio that will give the desired coupling with both grid and plate connected to the end of the tank circuit.

Coupling Condensers

Coupling condensers should be of the mica type with a voltage rating above the sum of the driver plate and amplifier biasing voltages.

Amplifier Design Considerations

● PLATE-CIRCUIT VALUES

Tank-Circuit Q

Power cannot be readily coupled out of a plate tank circuit if the Q is too low. Also, harmonics are more readily coupled out of a tank circuit whose Q is low. On the other hand, a large C/L ratio causes high circulating current in the tank circuit, increasing the losses. Unless one of these factors is considered to be of greater importance than the other, a compromise Q value of 12 usually is selected.

With the conditions under which r.f. power amplifiers in amateur transmitters usually are operated, the L/C ratio for the same Q varies in proportion to the ratio of d.c. plate voltage to plate current with the amplifier in operation and loaded. The chart of Fig. 6-22 shows recommended values of tank capacitance for a Q of 12 for a wide range of plate-voltage/plate-current ratios for each of the low-frequency amateur bands. The values given apply to the type of plate tank circuits shown in Fig. 6-23A and B only. Because the tube is connected across only half of the tank in the remainder of the circuits shown in Fig. 6-23, the total capacitance across the tank coil may be reduced to one-quarter that shown by the graph for the same plate-voltage/plate-current ratio. This means that in circuits in which a split-stator condenser is used, the capacitance of each section of the condenser may be half the value shown in the graph, since the two sections are in series across the coil.

The values shown in Fig. 6-22 are the capacitances which should be in actual use when the circuit is tuned to resonance in the selected band — not the maximum rated capacitance of the tank condenser — including tube and circuit capacitances. They should be considered minimum values for satisfactory operation. They can be exceeded 50 to 100 per cent without involving an appreciable loss in circuit efficiency. The Q can be increased also by tapping the plate down on the tank coil, although this sometimes results in setting up a parasitic oscillatory circuit.

Plate Tank-Condenser Voltage

In selecting a tank condenser with a spacing between plates sufficient to prevent voltage breakdown, the peak r.f. voltage across a tank circuit under load, but without modulation, may be taken conservatively as equal to the d.c. plate voltage. If the d.c. plate voltage also appears across the tank condenser, this must be added to the peak r.f. voltage, making the total peak voltage twice the d.c. plate voltage. If the amplifier is to be plate-modulated, this last value must be doubled to make it four times the d.c. plate voltage, because both d.c. and r.f. voltages double with 100-per-cent plate modulation. At the higher plate voltages, it is desirable to choose a tank circuit in which the d.c. and modulation voltages do not appear across the tank condenser, to permit the use of a smaller condenser with less plate spacing. Fig. 6-23 shows the peak voltage, in terms of d.c. plate voltage, to be expected across the tank condenser in various circuit arrangements. These peak-voltage values are given assuming that the amplifier is loaded to rated plate current. Without load, the peak r.f. voltage will run much higher. Since a c.w. transmitter may be operated without load while adjustments are being made, although a modulated amplifier never should be operated without load, it is sometimes considered logical to select a condenser for a c.w. transmitter with a peak-voltage rating equal to that required for a 'phone transmitter of the same power. However, if minimum cost and space are considerations, a condenser with half the spacing required for 'phone operation can be used in a c.w. transmitter for the same carrier output, as indicated under Fig. 6-23, if power is reduced temporarily while tuning up without load.

In the circuits of Fig. 6-23C, D and E, the rotors are deliberately connected to the positive side of the high-voltage supply, eliminating any difference in d.c. potential between the rotors and stators.

The plate spacing to be used for a given peak voltage will depend upon the design of the variable condenser, influencing factors being the mechanical construction of the unit, the dielectric used and its placement in respect to intense fields, and the condenser-plate shape and degree of polish. Condenser manufacturers usually rate their products in terms of the peak voltage between plates.

Plate Tank Coils

The inductance of a manufactured coil usually is based upon the highest plate-voltage/plate-current ratio likely to be used at the maximum power level for which the coil is designed, following the logical conclusion that it is easier to cut off turns than to add them. Therefore in the majority of cases, the capacitance shown by Fig. 6-22 will be greater than

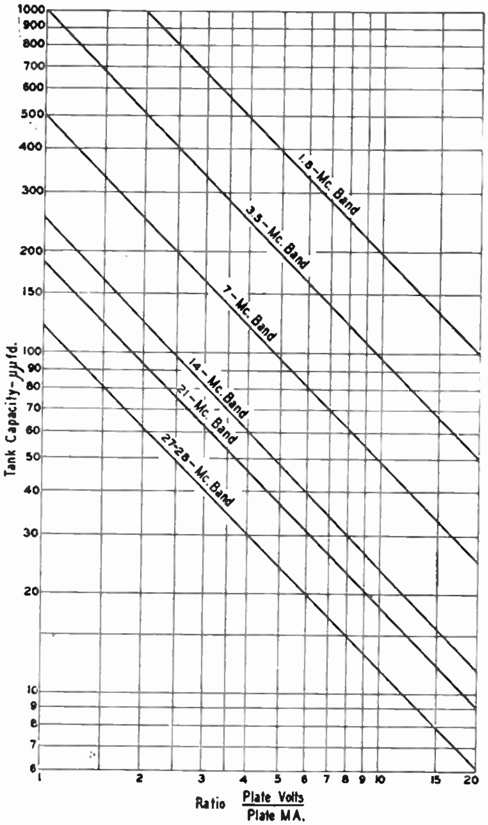


Fig. 6-22 — Chart showing minimum plate tank capacitances recommended with various ratios of plate voltage to plate current, for the six low-frequency amateur bands. In the circuits F, G and H of Fig. 6-23, the values shown by the graph may be divided by four. In circuits C, D, E, I, J and K, the capacitance of each section of the split-stator condenser may be one-half the value shown by the graph. The full graph values should be used for circuits A and B. These values are based on a circuit Q of 12.

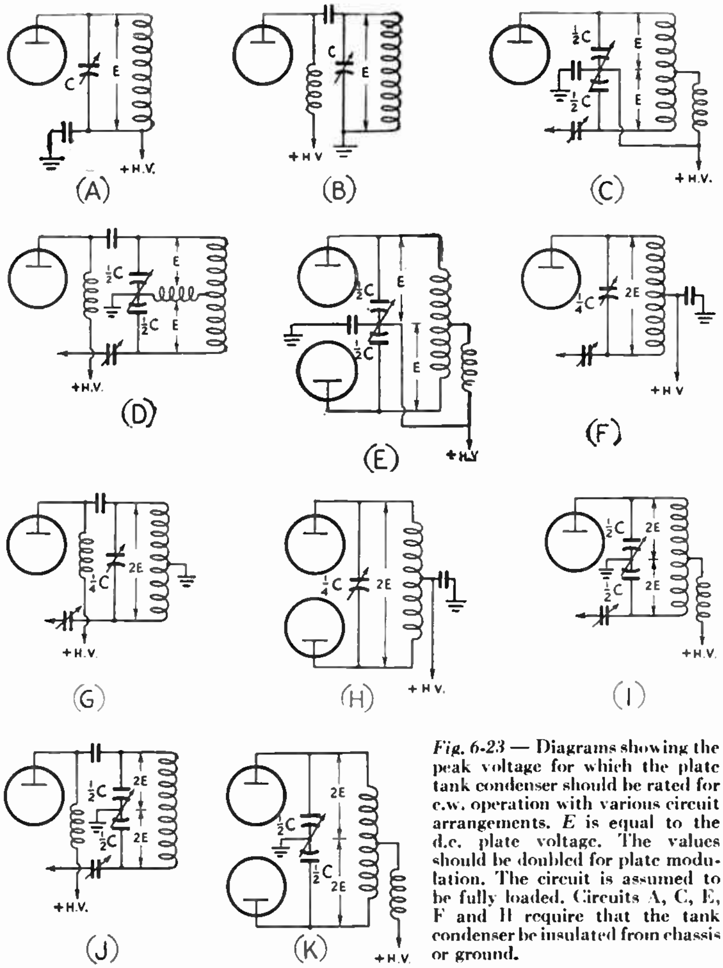


Fig. 6-23 — **Diagrams showing the peak voltage for which the plate tank condenser should be rated for c.w. operation with various circuit arrangements. *E* is equal to the d.c. plate voltage. The values should be doubled for plate modulation. The circuit is assumed to be fully loaded. Circuits A, C, E, F and H require that the tank condenser be insulated from chassis or ground.**

that for which the coil is designed and turns must be removed to permit the use of the proper value of capacitance. At 28 Mc., and sometimes 14 Mc., the value of capacitance shown by the chart for a high plate-voltage/plate-current ratio will be lower than that attainable in practice with the components available. The design of manufactured coils usually takes this into consideration also and it may be found that values of capacitance greater than those shown in the graph (if stray capacitance is included) are required to tune these coils to the band.

Manufactured coils are rated according to the plate power input to the tube or tubes when the stage is loaded. Since the circulating tank current is much greater when the amplifier is unloaded, care should be taken to operate the amplifier conservatively when unloaded to prevent damage to the coil as a result of excessive heating.

Plate-Blocking and By-Pass Condensers

Plate-blocking condensers should have low inductance; therefore condensers of the mica type are preferred. The capacitance should be large enough to have low reactance at the lowest operating frequency. For frequencies between 3.5 and 30 Mc., a capacitance of 0.001 μfd. is commonly used. The voltage rating should be 25 to 50 per cent above the plate-supply voltage.

By-pass condensers also should have low reactance at the operating frequency. Paper condensers with a capacitance of 0.01 μfd. are satisfactory for supply voltages up to 500 or 600 at frequencies up to at least 7 Mc. Mica condensers, usually 0.001 μfd., are preferable at the higher frequencies and greater plate voltages.

Voltage ratings should be doubled in the case of plate modulation.

R.F. Chokes

Parallel plate feed provides a considerable measure of protection against serious injury to the operator from accidental contact with high-voltage d.c. in the tank circuit. However, the r.f. choke in this case is called upon to present a high impedance at the operating frequency if serious loss of power in the choke is to be avoided. In the design of manufactured r.f. chokes, an attempt is made to make the choke universally satisfactory for several amateur bands. However, when the transmitter is designed to operate on all amateur bands from 28 Mc. to 3.5 Mc., loss in r.f. chokes often occurs on one or more of the bands. There is no simple remedy for this difficulty aside from a shift to series plate feed which, of course, nullifies the safety angle. One possible remedy is the use of different chokes for each band, the chokes being plugged in with the tank coil.

For frequencies between 3.5 and 30 Mc., 2.5-mh. chokes are used where the plate current is 125 ma. or less, and 1 mh. when the plate current is above 125 ma. In the circuit of Fig. 6-23D, the choke does not carry any current, so a low-current choke may be used, regardless of the power. In series-fed circuits in which the choke is used to isolate the coil center-tap from ground, the value of the choke inductance is not critical.

GRID TANK CIRCUITS

The value of capacitance to be used in a grid tank circuit when employing link coupling is not critical so long as the Q is high enough to permit satisfactory coupling to the driver stage. A capacitance of 200 $\mu\mu$fd. should be sufficient in most cases for unbalanced grid tank circuits tuned to 3.5 Mc., with the value decreased in proportion as the frequency increases, as given under Fig. 6-24. For unbalanced grid tank circuits, the total condenser capacitance may be cut in half, making the capacitance of each section of a split-stator condenser the same as that of the single condenser used in an unbalanced input grid tank circuit.

The Q can also be increased by tapping the grid down on the input coil, at some risk, however, of setting up a parasitic circuit.

Approximate tank-condenser voltage ratings are suggested under Fig. 6-24. Tank coils with a power rating equal to that of the driver plate tank coil should be used in the grid tank circuit.

The resistor R in Fig. 6-24C and D is recommended in place of the r.f. choke customarily used in the same position, to eliminate the possibility of forming a low-frequency parasitic t.g.t.p. oscillator in conjunction with the r.f. choke usually used similarly in the plate circuit. A resistance of 100 ohms will be sufficient in most cases. If a grid leak is used, the 100-ohm resistor will not be necessary.

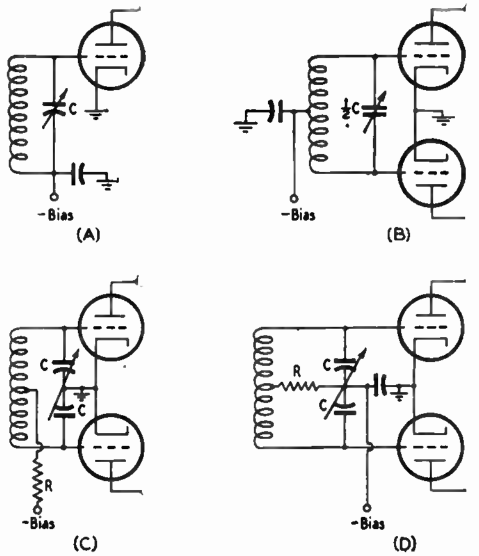


Fig. 6-24 — Diagrams for determining grid tank-condenser capacitance. C should be approximately 200 $\mu\mu$fd. for 3.5 Mc., 100 $\mu\mu$fd. for 7 Mc., 50 $\mu\mu$fd. for 14 Mc. and 25 $\mu\mu$fd. for 28 Mc.

The tank condenser should have a voltage rating approximately equal to the operating bias voltage plus 20 per cent of the plate voltage for circuit A, twice this value for circuit B and each section of the condenser in circuit D, while the biasing voltage must be added to this latter figure in determining the voltage rating of each section of the condenser in circuit C. R is an isolating resistor of 100 ohms.

R. F. Power-Amplifier-Tube Operating Factors

Transmitting-tube instruction sheets and data tables specify the limitations on various electrode voltages and currents which should be observed in the operation of transmitting tubes to insure normal tube life. Included also are sets of recommended operating conditions which may be followed as a guide in obtaining rated output with good efficiency consistent with reasonable driving power, although it may be desirable to depart from these somewhat under certain conditions.

GRID-CIRCUIT RATINGS

Grid Bias

Two values of grid-biasing voltage are of interest in the practical operation of r.f. power amplifiers. These are **protective bias** and **operating bias.**

When plate (and screen) voltage is applied, most tubes will draw appreciable plate current in the absence of any grid bias. Therefore protective bias must be used with all but "zero-bias"-type tubes to hold the power input to the tube below the rated dissipation value when excitation is removed without removing plate (and screen) voltage. Without excitation, the amplifier delivers no power. Therefore any power input is dissipated in heat which would ruin the tube in a short length of time. This condition exists when the transmitter is keyed ahead of the amplifier, while tuning adjustments are being made, or through failure of a crystal oscillator to function or other accidental failures.

Operating bias is the value of biasing voltage between grid and cathode when the amplifier is being driven and delivering power. The optimum value of biasing voltage for operating under a given set of conditions is listed in tube tables and manuals, and with triodes is normally two to three times the **cut-off bias value** — the value necessary to reduce the plate current to zero with plate voltage applied.

Protective bias may be any value between that which limits the input to the tube to its rated plate (and screen) dissipation as a minimum, and the operating value as a maximum. It is common practice, however, to set the value at some point between that which is necessary to cut off plate current completely and the operating value. With fixed plate voltage, the cut-off value for a triode can be determined quite closely by dividing the plate voltage by the amplification factor obtained from the tube data sheet. For screen-grid tubes,

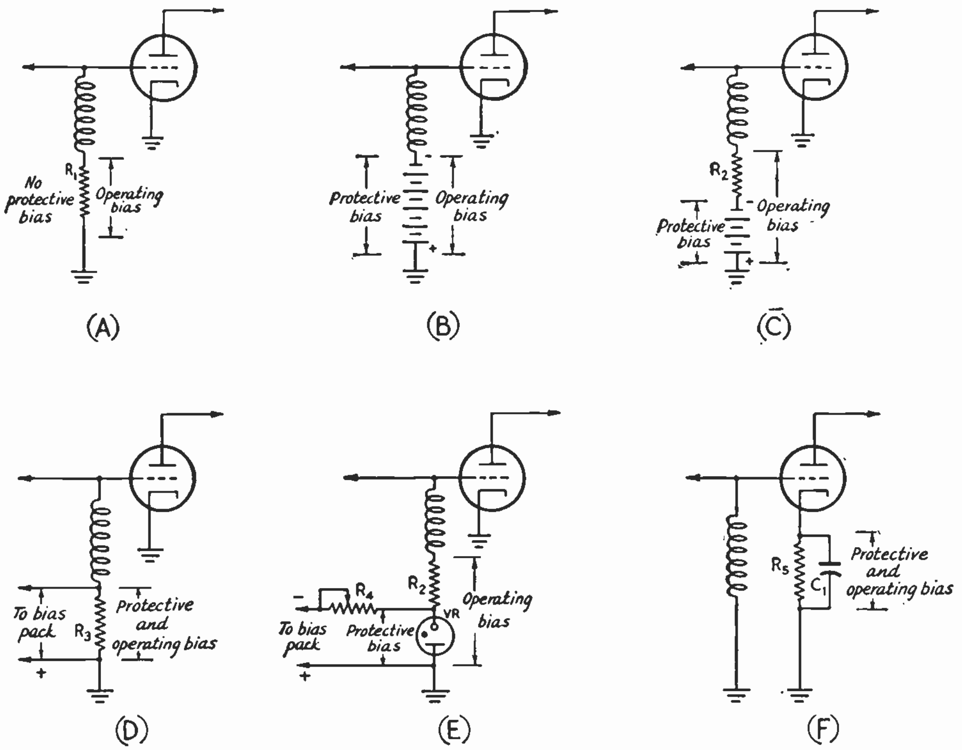


Fig. 6-25 — **Various systems for obtaining protective and operating bias for r.f. amplifiers. A — Grid-leak. B — Battery. C — Combination battery and grid leak. D — Grid leak and adjusted-voltage bias pack. E — Combination grid leak and voltage-regulated pack. F — Cathode bias.**

the amplification factor and voltage of the screen must be used instead. In cases where this is not included in the operating data, the approximate cut-off value may be obtained from an inspection of the plate-current plate-voltage curves which show the plate current for a wide range of plate, screen and biasing voltages.

A saving in the operation of a c.w. amplifier sometimes can be effected by adjusting the *protective* bias so that the tube (or tubes if more are operated from the same supply) draws the same current as the required bleeder resistance for the power supply (see Chapter Seven), if this can be done without exceeding the dissipation rating of the tube. This saves the cost of the bleeder resistor and some of the power it wastes and also improves the regulation, since the difference between minimum and maximum load as the amplifier is keyed is less.

A factor which must be considered in determining the value of bias which will protect the tube is plate- (and screen-) voltage regulation. If the power-supply regulation is poor, or if the plate or screen is fed from a resistance voltage divider or a voltage-dropping resistor, the electrode voltages will soar as the tube draws less than normal operating current and therefore an increase over the calculated value of cut-off bias will be required to bring the current to zero. This condition is encountered most often in the operation of a screen-grid tube where the screen is not fed from a fixed-voltage source. In such cases, care should be taken to make certain that the proper operating bias is not exceeded when excitation is applied.

Several different systems for obtaining bias are shown in Fig. 6-25. At A, bias is obtained entirely from the voltage drop across the grid leak, R_1, caused by the flow of rectified grid current when the amplifier is being driven. This system has the desirable feature that the biasing voltage, being dependent upon the value of grid current, is kept adjusted close to proper operating value automatically over a considerable range of excitation levels. However, when excitation is removed, grid-current flow ceases and the voltage across R_1 falls to zero and there is no bias. Therefore this system provides no protection for the amplifier tube in case excitation fails or is removed.

A battery delivering the required operating bias is used in the arrangement of Fig. 6-25B. Since the biasing voltage still remains when excitation is removed, plate-current flow ceases and the tube is protected. A factor which must be taken into consideration when dry batteries, such as "B" batteries, are used, is the resistance of the batteries. If the internal resistance is high, the resistance will cause an increase.

by grid-leak action, in the operating bias above that normally delivered by the batteries. Batteries develop internal resistance with age and should be replaced from time to time. Another factor is that the direction of grid-current flow is such as to reverse the normal direction of current through the battery. This acts to charge the battery. A battery which has been in use for some time, particularly if the grid current under excitation is high, will show a considerably higher-than-rated terminal voltage because of the charging action of the grid current. The terminal voltage of a battery used in transmitter bias service where grid current flows cannot be used as an indication of the condition of the battery. Its internal resistance may be high, even though it shows normal or above-normal terminal voltage. If the grid current in a battery-biased stage falls off after a period of operation and no other reason is obvious, it is probable that the biasing battery should be replaced. The battery life which may be expected in bias service with a given value of grid current will be approximately the same as it would be if that same current were being drawn from the battery.

In Fig. 6-25C, the battery voltage is reduced to the protective value. When excitation is applied, grid-leak action through R_2 supplies the additional biasing voltage necessary to bring the total up to the operating value. This combination of fixed and grid-leak bias is the most popular system, since it combines the safety of protective fixed bias and a measure of automatic adjustment of the operating value through grid-leak action.

In Fig. 6-25D, a power pack is used to supply protective bias. The output of the power pack is connected across the grid resistor which is of the normal grid-leak value for the tube. The peak voltage output of the transformer used in the power pack must not exceed the operating-bias value. A bleeder resistance cannot be used across the output of the pack, nor can the output voltage be reduced by means of a voltage divider or series dropping resistor without affecting the biasing voltage when excitation is applied.

These restrictions on the use of a power pack can be avoided by the addition of a voltage-regulator tube across the output of the pack, as shown in Fig. 6-25E. The voltage across the regulator tube remains constant with or without grid current flowing. By making the voltage-regulator series resistor, R_1, of proper value, the output voltage of the pack may be anything within reason above a minimum of approximately twice the voltage rating of the VR tube. These tubes are available for 75, 90, 105 and 150 volts and each tube will handle up to 30 or 40 ma. of grid current. VR tubes may be used in series to obtain regulated voltages above 150, and in parallel for grid currents above 40 ma. It is usual practice to use a VR tube, or combination of VR tubes in series or series-parallel, with the minimum voltage rating which will give plate-current cut-off, and obtain the additional voltage required to bring the total bias up to the operating value by grid-leak action when excitation is applied, as with battery bias in Fig. 6-25C. The use of VR tubes for this purpose is discussed more fully in Chapter Seven.

A single source of fixed biasing voltage, such as batteries or VR tubes in series, may be used to provide protective bias for more than one amplifier stage, tapping the batteries or connecting to the junction of the tubes in the VR series if lower biasing voltages are required for other stages. In this case, the current flowing through the fixed-bias source is the sum of the grid currents of the individual stages obtaining bias from the source.

In Fig. 6-25F, bias is obtained from the voltage drop across a resistor in the cathode (or filament center-tap) lead. Protective bias is obtained by the voltage drop across R_5 as a result of plate (and screen) current flow. Since plate current must flow to obtain a voltage drop across the resistor, it is obvious that cut-off protective bias cannot be obtained by this system. When excitation is applied, plate (and screen) current increases and the grid current also contributes to the drop across R_5, thereby increasing the bias to the operating value. Since the voltage between plate and cathode is reduced by the amount of the voltage drop across R_5, the over-all supply voltage must be the sum of the plate and operating-bias voltages.

The resistance of R_5 should be adjusted to the value which will give the correct operating bias with rated grid, plate and screen currents flowing with the amplifier loaded to rated input. When excitation is removed, the input to most types of tubes will fall to a value that will prevent damage to the tube, at least for the period of time required to remove plate voltage.

Calculating Bias-Resistor Values

The calculation of the required grid-leak and cathode biasing-resistor values is not difficult. For simple grid-leak bias, as shown in Fig. 6-25A, the resistance is obtained by dividing the required operating-bias voltage by the rated grid current.

Example: Required operating bias = 100 volts.
Rated grid current = 20 ma. = 0.02 amp.

$$\text{Grid-leak resistance} = \frac{100}{0.02} = 5000 \text{ ohms.}$$

If a combination of grid-leak and fixed protective bias is used, the amount of protective bias should be subtracted from the required operating-bias voltage before the calculation is made (except in the case of the arrangement of Fig. 6-25D).

Example: Required operating bias = 150 volts.
Protective bias from battery or VR tube = 90 volts.
150—90 = 60 volts = required bias from grid leak.
Rated grid current = 10 ma. = 0.01 amp.

$$\text{Grid-leak resistance} = \frac{60}{0.01} = 6000 \text{ ohms.}$$

In the case of a cathode biasing resistor, the rated grid, screen and plate currents under load are added together. The required operating voltage is then divided by this total current to obtain the resistance.

Example: Rated grid current = 15 ma. = 0.015 amp.
Rated screen current = 20 ma. = 0.02 amp.
Rated plate current = 200 ma. = 0.2 amp.
Total rated cathode current = 235 ma. = 0.235 amp.
Required operating bias = 150 volts.

$$\text{Cathode resistance} = \frac{150}{0.235} = 638 \text{ ohms.}$$

For two tubes in parallel or push-pull that use a single common resistor in examples similar to those above, the calculated value of resistance should be cut in half.

The power rating of the resistor may be determined from Ohm's Law:

$$P = I^2R$$

Example: In the first example above for grid-leak resistance,
I = 20 ma. = 0.02 amp. I^2 = 0.0004
R = 5000 ohms.
P = (0.0004) (5000) = 2 watts.

Example: In the above example for cathode resistor,
I = 235 ma. = 0.235 amp. I^2 = 0.055
R = 638
P = (0.055) (638) = 35.1 watts.

Maximum Grid Current

When a Class C amplifier is properly excited, and the grid is driven positive over part of the cycle, rectification takes place as it does in a diode. The rectified grid current flows between grid and cathode within the tube and thence through the external d.c. circuit which must always be provided, connecting grid and cathode. This external circuit includes the bias source (grid leak or voltage source) and either the grid r.f. choke with parallel feed, or the tank coil in series-feed arrangements. The flow of rectified current causes heating of the grid. As with the plate, there is a limit to the heat which the grid can dissipate safely. This limit is expressed in terms of maximum d.c. grid current which should not be exceeded in regular operation of the amplifier. Efficient operation usually can be attained with grid current below the maximum rated value.

The rated total grid current of two tubes in parallel or push-pull is twice that of a single tube of the same type.

Excitation

Excitation, or driving power, is the r.f. power fed to the grid of the amplifier by a preceding oscillator or amplifier. For efficient operation, a triode amplifier requires a driver capable of delivering 15 to 20 per cent as much power as the rated output of the amplifier. Screen-grid tubes require much less — usually from 5 to 10 per cent of their rated power output. To cover tank-circuit and coupling losses, a driver capable of supplying several times the driving power listed in the tube data should be used.

Two tubes in parallel or push-pull require twice the driving power of a single tube of the same type under similar conditions.

PLATE-CIRCUIT RATINGS

Power Output

The figure for power output given in the tube data is the r.f. power that the tube can be expected to deliver to the tank circuit (not the power output from the tank which is somewhat lower) under the conditions specified, at the fundamental frequency.

Power Input

Power input for both triodes and screen-grid tubes is the d.c. power input to the plate circuit. It is the product of the d.c. plate voltage and plate current.

Example: Plate voltage = 1250 volts.
Plate current = 150 ma. = 0.15 amp.
Power input = (1250) (0.15) = 187.5 watts.

Plate and Screen Dissipation

All of the d.c. power fed to the plate circuit of an amplifier is not converted into r.f. power. Part of it is wasted in heat within the tube. There is a limit to the amount of power that a tube can dissipate in the form of heat without danger of damage to the tube. This is the maximum rated plate dissipation given in tube data. The power dissipated is the difference between the d.c. power input and the r.f. power output.

Since the d.c. power furnished to the screen of a pentode or tetrode does not contribute to the r.f. output, it is entirely dissipated in heating the screen, and the maximum-input rating should be carefully observed.

Plate Efficiency

The efficiency of an amplifier is the ratio of r.f. power output to the d.c. power input.

Example: D.c. power input = 175 watts.
R.f. power output = 125 watts.
Dissipation = 175—125 = 50 watts.

$$\text{Efficiency} = \frac{125}{175} = 0.714 = 71.4 \text{ per cent.}$$

The plate efficiency at which an r.f. power amplifier can be operated depends chiefly upon

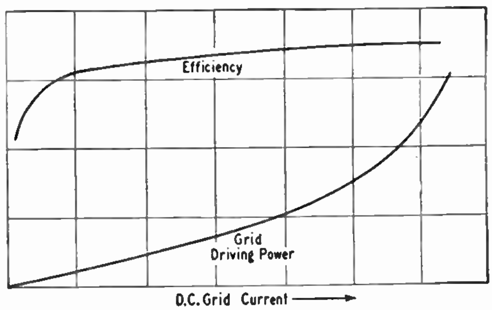


Fig. 6-26 — Curve showing relation between driving power and plate-circuit efficiency of an r.f. power-amplifier stage.

the relative driving power delivered to the input circuit. Fig. 6-26 shows that the driving power must be increased considerably out of proportion to the increase in efficiency at the higher efficiencies. An efficiency of 65 to 75 per cent represents a satisfactory balance between power output and driving power.

Maximum Plate Current and Voltage

All voltage figures given in tube data, unless otherwise specified, refer to the voltage between the electrode mentioned and cathode, or filament center-tap. Included are figures for maximum rated plate voltage and plate current. These are the respective maximum values that should be used under any circumstances. Neither should be exceeded to compensate for a lower-than-maximum value of the other in attempting to bring the power input up to permissible level. These maximum values should not be used simultaneously unless it is possible to do so without exceeding the rated plate dissipation.

OTHER OPERATING FACTORS

Filament Voltage

The filament voltage for the indirectly-heated cathode-type tubes found in low-power classifications may vary 10 per cent above or below rating without seriously reducing the life of the tube. But the voltage of the higher-power filament-type tubes should be held closely between the rated voltage as a minimum and 5 per cent above rating as a maximum. Care should be taken to make sure that the plate power drawn from the power line does not cause a drop in filament voltage below the proper value when plate power is applied. When the filament transformer is found not to deliver the required filament voltage, the voltage may be adjusted by means of a resistor in series with the transformer primary if the transformer voltage is too high, or by one of the line-voltage adjusting schemes described in Chapter Seven that either boosts the voltage or reduces it as necessary.

Thoriated-type filaments lose emission when the tube is overloaded appreciably. If the overload has not been too prolonged, emission sometimes may be restored by operating the filament at rated voltage with all other voltages removed for a period of 10 minutes, or at 20 per cent above rated voltage for a few minutes.

Interelectrode Capacitances

The value given in tube data for grid-plate capacitance is useful in determining the value of capacitance necessary to neutralize a triode. The input- and output-capacity values are helpful in arriving at a figure of minimum circuit capacitance, particularly where capacitive coupling is used.

Adjustment of R. F. Amplifiers

GENERAL TUNING PROCEDURE

Metering

Sets of typical operating conditions for r.f. amplifiers are given in all tube-data sheets and these should be followed closely for maximum output with a good balance between efficiency and required driving power. In amateur service, ICAS (intermittent commercial-amateur service) ratings may be used when this set of ratings is given. When the available plate voltage falls between values given in the data, satisfactory performance may be obtained by using intermediate values for the other voltages and currents listed.

Fig. 6-27 shows the connections for a voltmeter and milliammeter to obtain desired readings. While cathode metering often is used for reasons of safety to the operator and meter insulation, it is frequently difficult to interpret readings that are the resultant of three currents, one of which may be falling while the other two are increasing. Fig. 6-28 shows a commonly-used system for switching a single meter to read current in any of several different circuits. The resistors, R, are connected in the various circuits in place of the milliammeters shown in Fig. 6-27. Since the resistance of R is several times the internal resistance of the milliammeter, it will have no practical effect upon the reading of the meter itself.

When the meter must read currents of widely differing values, a meter with a range sufficiently low to accommodate the lowest values of current to be measured may be selected. In the circuits in which the current will be above the scale of the meter, the resistance of R can be adjusted to a lower value which will give the meter reading a multiplying factor. (See Chapter Sixteen.) Care should be taken to observe proper polarity in making the connections between the resistors and the switch.

Input-Circuit Adjustment

In setting up an r.f. power amplifier for operation, the necessary provisions for grid bias should be made first. ("R.F. Power-Amplifier Tube Operating Factors," this chapter.) The output of the driver (the oscillator and whatever intermediate amplifier stages there may be) should have been checked previously and found to be adequate. The amplifier biasing system should be connected, and if it includes a fixed protective supply, this should be turned on. No plate or screen voltage should be applied to the amplifier, however.

In general, with capacitive coupling, an amplifier grid-current reading should be obtained when the driver is coupled to the amplifier and tuned to resonance. If the driver is

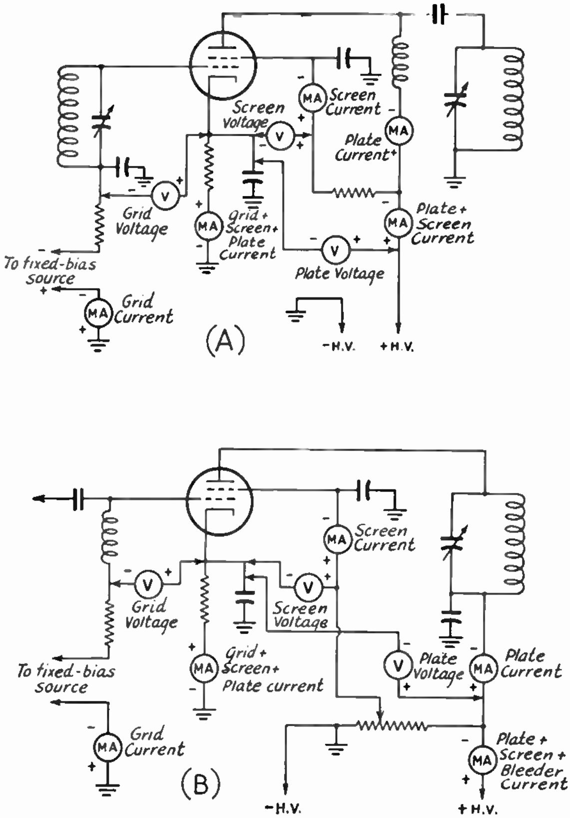


Fig. 6-27 — Diagrams showing placement of voltmeter and milliammeter to obtain desired measurements. A — Series grid feed, parallel plate feed and series screen voltage-dropping resistor. B — Parallel grid feed, series plate feed and screen voltage divider.

a simple VFO or a crystal oscillator of the Pierce type, with no separate tuned output-circuit tank, the operation is merely one of adjusting the coupling to the amplifier until rated amplifier grid current, or the maximum consistent with satisfactory oscillator stability, is obtained. If link coupling is used, the grid tank circuit must also be tuned to resonance as indicated by the peak in grid current.

With all capacitive-coupled drivers having a tuned output tank, maximum amplifier grid current should occur at or very close to the point where the driver plate current dips to a minimum. With link coupling, the amplifier grid tank condenser should first be set at minimum or maximum, whichever is judged to be farthest from resonance. The driver output circuit should then be tuned for minimum plate current. Then the grid tank condenser should be swung for maximum grid current. As a final tuning adjustment, the driver plate tank circuit should be retuned to make sure that it is at the minimum point of its plate-current dip. As the coupling is increased, the driver plate-current dip will become less pronounced and may almost disappear altogether if the coupling is increased sufficiently. This, however, usually is an indication of driver overload. Maximum driver output (maximum amplifier grid-current reading) usually will be obtained with the coupling adjusted to the point where there is still a fair amount of dip in plate current. The dip is likely to be less with a fully-loaded screen-grid tube than with a triode. Each time an adjustment in coupling is made, the above tuning process should be repeated.

Proper excitation to an amplifier is indicated when the recommended grid current is obtained simultaneously with recommended grid bias, with the amplifier operating and fully loaded. But here, for preliminary tuning, any grid-current reading approximating the recommended value will suffice.

Output-Circuit Adjustment

At this point, the driver should be turned off and the amplifier checked for parasitic oscillation. (See "Parasitic Oscillations," this chapter.)

The next step in the adjustment of a triode amplifier is that of neutralization. (See "Neutralizing Procedure," this chapter.)

After the amplifier stage has been stabilized, the output circuit may be adjusted. With normal bias and excitation applied again, reduced plate voltage can now be turned on and the plate tank circuit resonated.

Resonance in the plate circuit of an r.f. power amplifier is accompanied by a dip in plate current similar to that shown in Fig. 6-11. This dip is caused by the increase in tank impedance in the plate circuit when the tank is tuned to resonance. When the tank is not at resonance, the plate-circuit impedance is low and therefore the plate current is high. An external load coupled to the tank circuit lowers the impedance and therefore the plate current at resonance increases.

If no other means is available for reducing plate voltage, a 115-volt lamp of 100- to 150-watt size may be connected in series with the primary of the plate transformer, provided it is separate from the transformer supplying filaments. A dummy load (see "Checking Power Output," this chapter) should now be coupled to the output tank circuit and the tank retuned to resonance. The minimum plate current at the dip at resonance should be higher after the load is connected and the dummy load should show an indication of output. Full plate voltage may now be applied and the plate tuning checked carefully for the dip at resonance. When testing at full plate voltage, care should be taken not to operate the amplifier off resonance longer than absolutely necessary, because the tube may be seriously damaged.

If the plate current at full voltage is not up to the rated value, the coupling to the load should be increased until the plate current at resonance is the rated value. Under no circumstances should the plate circuit be detuned to obtain the desired increase in plate current, since this results in a decrease in power output and an increase in dissipation. If the plate current exceeds the rated value at resonance, the coupling to the load should be reduced.

Final Adjustment

The grid current and biasing voltage now should be checked while the amplifier is in operation under load. In a properly-neutralized triode amplifier, the grid current normally will fall off when plate voltage and load are applied. If it does not, it is an indication of regeneration and the amplifier should be checked for feed-back, either through the tube because of incomplete neutralization, or through paths external to the tube.

If the grid current falls below the recommended value when plate voltage and load are applied, the biasing voltage should be checked. If this is found to be above the recommended value, it should be decreased. This decrease in bias should serve to increase the grid current. If the grid current is still too low, or if the biasing voltage also checks low, the excitation must be increased by tightening the coupling to the driver or raising its plate voltage if either or both can be done without exceeding the driver-tube rating.

If the increase in excitation causes an increase in plate current to above the rated value, the coupling to the load should be reduced. The amplifier is correctly adjusted when all of the recommended values are obtained simultaneously.

SPECIAL ADJUSTMENT OF PUSH-PULL AMPLIFIERS

Proper push-pull operation requires an accurate balance between the two sides of the circuit. Otherwise the dissipation will not be distributed evenly between the two tubes, one being overloaded if an attempt is made to operate the amplifier at full rating. Unbalance is indicated when the grid and/or plate currents are not equal and, if serious, is accompanied by a visible difference in the color of the tube plates. If interchanging the tubes does not change the unbalance, the circuit is not symmetrical electrically.

If the coil center-tap in split-stator tank circuits is sufficiently well-isolated from ground, the balance will depend upon the accuracy of capacitance balance in the tank condensers, the length of leads connecting the tubes to the condenser (including the return lead from rotor to filament) and the settings of the neutralizing condensers. Unbalance in the plate circuit will seldom influence the balance in the grid circuit, but the opposite may not be true. Small differences often may be taken care of by a re-adjustment of the neutralizing condensers, possibly to slightly unequal settings. Lengthening one or the other of the leads between the tubes and the tank condenser will alter the balance, particularly in the plate circuit. In extremes it may be necessary to place a trimmer across one section of the split-stator condenser.

If the coil center-tap is grounded, unbalance usually can be corrected by shifting the coil center-tap. *Both* condenser and coil should not be grounded simultaneously, since this may result in a condition where the resonance point for each tube comes at a different setting of the tank condenser.

OPERATION OF SCREEN-GRID AMPLIFIERS

Most of the foregoing procedure relating to triodes applies also to screen-grid tubes. However, principally because of the presence of the screen, there are additional factors which must be considered. Most screen-grid transmitting tubes are designed to operate without neutralization. However, this assumes certain further considerations. Because of the high power-sensitivity of such tubes, the feed-back coupling needed for oscillation is very small. Beyond the requirement of a well-screened tube, any possible feed-back coupling external to the tube must be reduced to a minimum. Special care must be used in the construction so that the input and output tank-circuit components and their respective wiring are well isolated from each other through judicious placement,

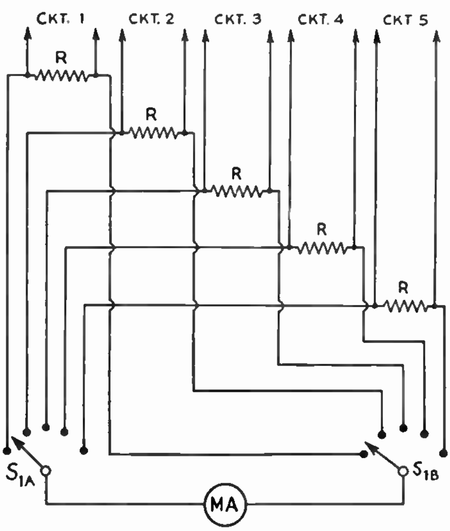


Fig. 6-28 — Method of switching a single milliammeter. The resistors, *R*, should be 10 to 20 times the internal resistance of the meter; 47 ohms will usually be satisfactory. S_1 is a 2-section rotary switch. Its insulation should be ceramic for high voltages, and an insulating coupling should always be used between shaft and control knob.

and by shielding as completely as possible. Because it is sometimes difficult to eliminate all external capacitive coupling, it may be necessary to neutralize a screen-grid amplifier to eliminate all tendency toward oscillation.

Considerable dependence must be placed also on the fact that, from other considerations, a screen-grid amplifier should always be operated fully loaded, since the loading helps to prevent oscillation. Return leads to cathode,

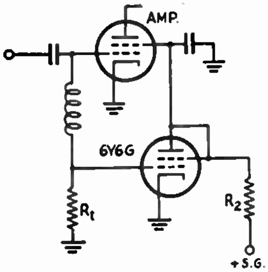


Fig. 6-29 — Screen protective circuit for screen-grid amplifier as an alternative to the use of fixed bias. R_1 is the normal grid leak for the amplifier and R_2 the recommended screen voltage-dropping resistor.

common to both plate and grid circuits, should be avoided. It is particularly important that the cathode be grounded directly or by-passed at the socket terminal and that the screen be by-passed thoroughly to the cathode with a mica condenser and short leads. The use of an un-by-passed parasitic-suppressing resistor at the screen is not recommended, since it aggravates instability at the operating frequency.

An indication of the coupling existing between input and output circuits can be obtained by the use of a sensitive r.f. indicator coupled to the output circuit as mentioned under "Neutralizing Procedure" in this chapter.

Other measures that can be taken to assist in stabilization at the operating frequency are the use of at least partial fixed bias and a non-resonant or detuned input circuit. With sufficient power from the driver, it is possible to secure rated excitation without having the grid circuit tuned close enough to resonance to start oscillation. In such a case the grid circuit should be detuned to the high-frequency side of resonance. Care should be taken that the grid circuit does not become resonant when the transmitter is tuned to another frequency.

Screen Considerations

For greatest protection to the tube, the screen voltage should be supplied from a series voltage-dropping resistor or a "light" voltage divider. When the screen is operated from a fixed-voltage source, the screen current increases rapidly with even slight amounts of overdrive or underloading. Since the series resistor serves to drop the voltage as the screen current increases, it affords a measure of protection. However, this same action may make it necessary to adjust the excitation with more than ordinary care if rated output is to be obtained. When a screen resistor or voltage divider is used, screen voltage should always be checked after each adjustment of excitation and loading to make sure that it is at rated value.

A screen-grid tube should never be operated at full screen voltage without plate voltage and full load. The screen current runs to damaging proportions under such conditions, especially if the screen is operated from a fixed-voltage source.

When plate and screen voltage and load are applied to a screen-grid amplifier, the grid current may increase, decrease or remain about the same, depending largely on the screen-voltage adjustment in relation to excitation.

Aside from the use of fixed bias, a screen-grid tube can be protected against excessive input when excitation is removed by the scheme shown in Fig. 6-29. A 6Y6G tetrode is connected as a low-μ triode. Since it is connected to the same point at the grid leak, the same bias appears at the grid of the protective tube and the grid of the amplifier. So long as excitation is supplied, the bias is sufficient to cut off the protective tube and it has no effect upon the operation of the amplifier. However, when excitation fails, the bias drops to zero and the 6Y6 draws current through the screen resistor, dropping the screen voltage to a point where the input to the amplifier is held within the dissipation rating.

CHECKING POWER OUTPUT

Dummy Loads

As a check on the operation of an amplifier, its power output may be measured by the use of a load of known resistance, coupled to the amplifier output as shown in Fig. 6-30. At A a thermoammeter, *A*, and a noninductive (ordi-

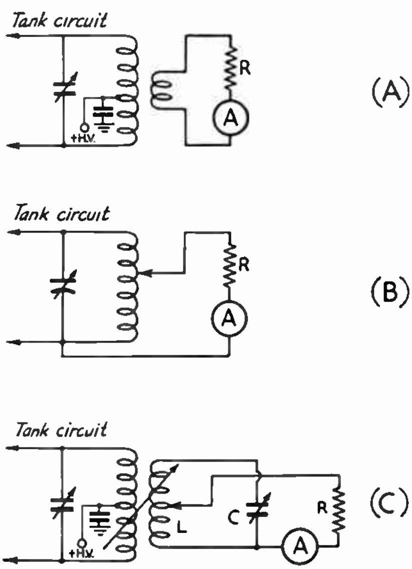


Fig. 6-30 — "Dummy-antenna" circuits for checking power output and making adjustments under load without applying power to the actual antenna.

nary wire-wound resistors are not satisfactory) resistance, R, are connected across a coil of a few turns coupled to the amplifier tank coil. The higher the resistance of R, the greater the number of turns required in the coupling coil. A resistor used in this way is generally called a **dummy antenna.** The loading may readily be adjusted by varying the coupling between the two coils, so that the amplifier draws rated plate current when tuned to resonance. The power output is then calculated from Ohm's Law:

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

where I is the current indicated by the thermoammeter and R is the resistance of the non-inductive resistor. Special resistance units are available for this purpose, ranging from 73 to 600 ohms (simulating antenna and transmission-line impedances) at power ratings up to 100 watts. For higher powers, the units may be connected in series-parallel. The meter scale required for any expected value of power output may also be determined from Ohm's Law:

$$I = \sqrt{\frac{P}{R}}$$

Incandescent light bulbs can be used to replace the special resistor and thermoammeter. The lamp should be equipped with a pair of leads, preferably soldered to the terminals on the lamp base. The coupling should be varied until the greatest brilliance is obtained for a given plate input. In using lamps as dummy antennas a size corresponding to the expected power output should be selected, so that the lamp will operate near its normal brilliancy. Then, when the adjustments have been completed, an approximation of the power output can be obtained by comparing the brightness of the lamp with the brightness of one of similar power rating in a 115-volt socket.

The circuit of Fig. 6-30B is for resistors or lamps of relatively high resistance. In using this circuit, care should be taken to avoid accidental contact with the plate tank when the power is on. This danger is avoided by circuit C, in which a separate tank circuit, LC, tuned to the operating frequency, is coupled to the plate tank circuit. The loading is adjusted by varying the number of turns across which the dummy antenna is connected on L and by changing the coupling between the two coils. With push-pull amplifiers, the dummy antenna should be tapped equally on either side of the center of the tank when the circuit of Fig. 6-30B is used.

Frequency Multiplication

SINGLE-TUBE MULTIPLIER

Output at a multiple of the frequency at which it is being driven may be obtained from an amplifier stage if the output circuit is tuned to a harmonic of the exciting frequency instead of to the fundamental. Thus, when the frequency at the grid is 3.5 Mc., output at 7 Mc., 10.5 Mc., 14 Mc., etc., may be obtained by tuning the plate tank circuit to one of these frequencies. The circuit otherwise remains the same as that for a straight amplifier, although some of the values and operating conditions may require change for maximum multiplier efficiency.

Efficiency in a single- or parallel-tube multiplier comparable with the efficiency obtainable when operating the same tube as a straight amplifier involves decreasing the operating angle in proportion to the increase in the order of frequency multiplication. Obtaining output comparable with that possible from the same tube as a straight amplifier involves greatly increasing the plate voltage. A practical limit as to efficiency and output within normal tube ratings is reached when the multiplier is operated at maximum permissible plate voltage and maximum permissible grid current. The plate current should be reduced as necessary to limit the dissipation to the rated value by increasing the bias. High efficiency in multipliers is not often required in practice, since the purpose is usually served if the frequency multiplication is obtained without an appreciable gain in power in the stage.

Since the input and output circuits are not tuned close to the same frequency, neutralization usually will not be required. Instances may be encountered with tubes of high transconductance, however, when a doubler will oscillate in t.g.t.p. fashion, requiring the introduction of neutralization. The link neutralizing system is convenient in such a contingency.

OTHER MULTIPLIER CIRCUITS

Push-Pull Multiplier

A single- or parallel-tube multiplier will deliver output at either even or odd multiples of

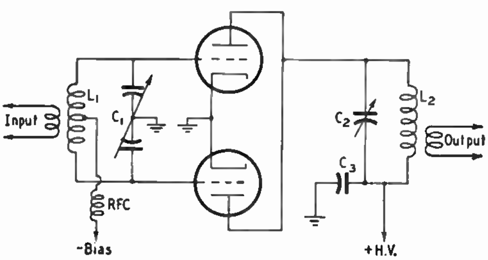


Fig. 6-31 — Circuit of a push-push frequency multiplier for even harmonics. The grid tank circuit, L_1C_1, is tuned to the frequency of the preceding driving stage, while the plate tank circuit, L_2C_2, is tuned to an even multiple of that frequency, usually the second harmonic. C_3 is the plate by-pass capacitor, usually a 0.01-μfd. paper condenser, while *RFC* is a 2.5-mh. r.f. choke.

the exciting frequency. A push-pull multiplier does not work satisfactorily at even multiples because even harmonics are largely canceled in the output. On the other hand, amplifiers of this type work well as triplers or at other odd harmonics. The operating requirements are similar to those for single-tube multipliers.

Push-Push Multipliers

A two-tube circuit which works well at even harmonics, but not at the fundamental or odd harmonics, is shown in Fig. 6-31. It is known as the **push-push** circuit. The grids are connected in push-pull while the plates are connected in parallel. The efficiency of a doubler using this circuit may approach that of a straight amplifier under similar operating conditions, because there is a plate-current pulse for each cycle of the output frequency.

This arrangement has an advantage in some applications. If the heater of one of the tubes is turned off, making the tube inoperative, its grid-plate capacitance, being the same as that of the remaining tube, serves to neutralize the circuit. Thus provision is made for either straight amplification at the fundamental with a single tube, or doubling frequency with two tubes as desired.

Multiplications of four or five sometimes are used to reach the bands above 28 Mc. from a lower-frequency crystal, but in the majority of lower-frequency transmitters, multiplication in a single stage is limited to a factor of two or three, because of the rapid decline in practicably obtainable efficiency as the multiplication factor is increased. Screen-grid tubes make the best frequency multipliers because their high power-sensitivity makes them easier to drive properly than triodes.

Parasitic Oscillations

Before placing the amplifier in operation, measures should be taken to make sure that the amplifier will function in a stable manner. In addition to the possibility of oscillation at or near the operating frequency, r.f. power amplifiers are subject to parasitic oscillation at frequencies far removed from the frequencies to which the amplifier is tuned by the conventional tank circuits. Oscillations of this type not only cause the transmission of illegal spurious signals, but they also impair the efficiency of the amplifier. In fact, they can be so severe as to make operation of the stage as an amplifier impossible and may destroy the tube if they are allowed to persist for any appreciable time. Erratic tuning characteristics invariably are a result of oscillation of one type or another. Parasitic oscillations may not be obvious under normal conditions of bias and load, but may be transient in nature, occurring intermittently during keying or modulation, causing widespread clicks or splatter. They can be treated most successfully only by adjusting the amplifier for conditions favorable toward sustained oscillation when they can be more readily observed and identified.

V.H.F. PARASITIC OSCILLATION

Parasitic oscillation in the v.h.f. range (usually in the vicinity of 100 to 200 Mc.) almost invariably will take place in an amplifier unless steps are taken to suppress it. Not always but in most cases, this sort of oscillation takes place as the result of an unavoidable t.g.t.p. circuit set up by the grid and plate leads tuned by the tank condensers in series, as shown by the heavy lines in Fig. 6-32A. The normal tank coils act only as r.f. chokes or capacitances at this high frequency. The same condition holds for balanced or push-pull circuits.

Testing Procedure

To test for this type of oscillation, the 28-Mc. tank coil should be plugged into the grid tank circuit (or the plate tank circuit of the driver stage if capacitance coupling is used) and the 3.5-Mc. coil in the plate tank circuit. This is to prevent any possible t.g.t.p. oscillation at the

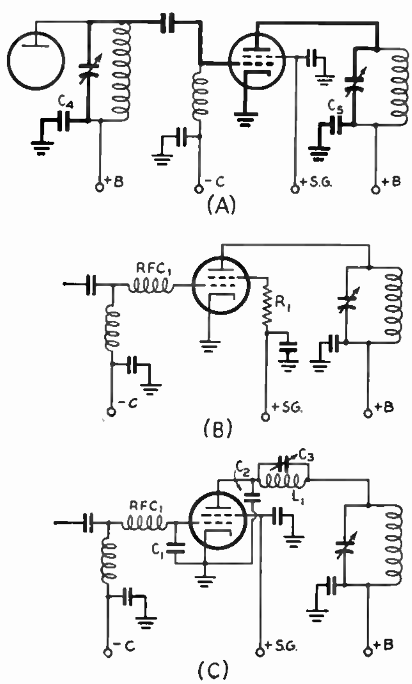


Fig. 6-32 — A — V.h.f. parasitic circuit hidden in high-frequency amplifier. B — One method of suppression with tetrodes. C — Preferred method. Approximate values: *RFC_1* — 15 turns No. 22, 1/4-inch diam., close-wound; *C_1* — 12-μμfd. ceramic; *C_2* — 15-μμfd. tubular; *C_3* — 100-μμfd. midget variable; *L_1* — 3 turns No. 14, 1/2-inch diam., 1/2 inch long.

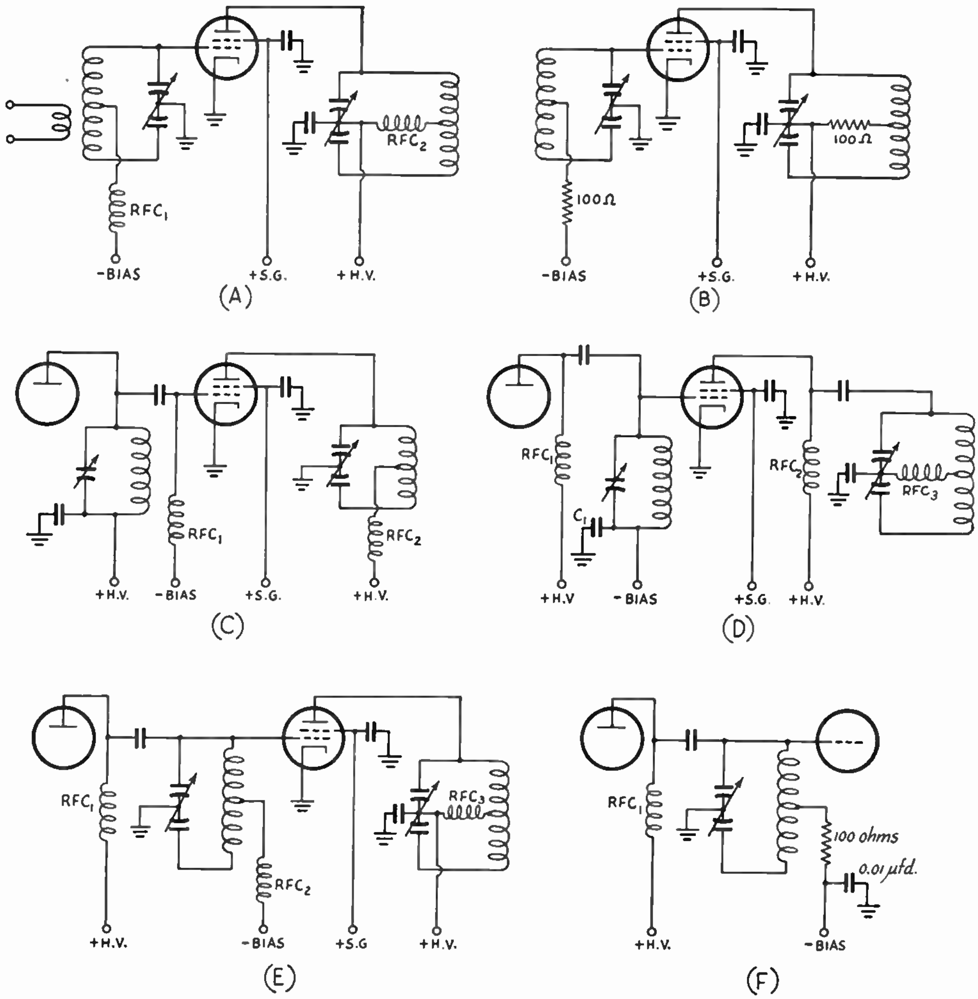


Fig. 6-33 — A, C and E show circuit arrangements to be avoided in eliminating low-frequency parasitic oscillation. The circuits of B, D and F are the recommended alternatives, which apply equally well to push-pull.

operating frequency which might lead to confusion in identifying the parasitic. If either tank circuit employs a split-stator condenser with an r.f. choke at the center of the coil, the r.f. choke should be short-circuited during the test. Any fixed bias should be replaced with a grid leak of 10,000 to 20,000 ohms. In a capacitive-coupled stage, the driver should be coupled in the normal way, but all load on the output of the amplifier should be disconnected. If the stage is an intermediate amplifier, the tube in the following stage should remain in place, but with its filament turned off. Plate (and screen) voltage should be reduced to the point where the rated dissipation is not exceeded. If a Variac is not available, voltage may be reduced by a 115-volt electric lamp of suitable wattage rating in series with the primary of the plate transformer.

With power applied only to the amplifier under test (not the driver), a careful search should be made by adjusting the input tank condenser to several settings, especially including minimum and maximum, and turning the plate tank condenser through its range for each of the grid-condenser settings. Any grid-current reading, or any dip or slight flicker in plate current at any point, indicates oscillation. This can be confirmed by using an indicating absorption wavemeter (see Chapter Sixteen) tuned to the frequency of the parasitic and held close to the plate lead of the tube.

Remedies

At the outset, an amplifier should be laid out so that the heavy leads shown in Fig. 6-32A are brought to the barest possible minimum. An inch of wire can be an appreciable length at 200 Mc. The inductance of those leads that cannot be made short can be reduced by the use of large conductor. Of equal importance are the return paths from the rotors of the input and output tank condensers to cathode or filament, usually made through the chassis.

The paths through the by-pass condensers (C_4 and C_5, Fig. 6-32A) to cathode should be as close to zero as possible. With capacitance coupling, this is often difficult, since the path through C_4 to *both* cathodes should be short. Link coupling has an advantage in this respect, since the grid return of the amplifier and the plate return of the driver are independent.

In the case of filament-type tubes, the filament should be by-passed directly at the socket with mica condensers and the returns made to the grounding point of the by-pass condensers.

V.h.f. parasitic oscillation usually can be suppressed in screen-grid tube circuits by inserting a v.h.f. choke in series with the grid and a small resistor of 50 to 100 ohms between the screen and its by-pass condenser, as shown in Fig. 6-32B. However, the introduction of even a small amount of resistance in the screen circuit in this manner invariably results in a reduction in the isolation between input and output circuits at the fundamental operating frequency. Therefore, unless the stage is to be neutralized, the treatment shown in Fig. 6-32C is preferable. Here, in addition to the v.h.f. choke at the grid, the input and output circuits are shunted by low-inductance condensers, C_1 and C_2. For amplifiers up to those requiring one or two 807s as drivers, small ceramic condensers connected across the grid and filament terminals of the tube sockets have been satisfactory. Those in the plate circuit should be of the tubular air type or vacuum type with a peak voltage rating equal to twice the plate voltage for c.w. operation, or four times the plate voltage for plate modulation. They should be connected with the shortest possible heavy leads between the plate cap at the top of the tube and the cathode or filament ground point. In the case of triode amplifiers, these condensers, when combined with short leads, often are the only requirement in eliminating v.h.f. parasitic oscillation.

In extreme cases of parasitic oscillation in screen-grid amplifiers, it may be necessary to add a v.h.f. wavetrap in the plate circuit as shown in Fig. 6-32C. The adjustment of such wavetraps in a push-pull amplifier will have a marked effect on the balance, however, and the two cannot be adjusted independently.

A sensitive grid-dip meter of the type described in Chapter Sixteen is often helpful in the locating of resonances responsible for parasitic oscillation. Once the circuit has been traced, it is easier to determine what can be done to detune or otherwise nullify the effects of the offending circuit.

LOW-FREQUENCY PARASITICS

Low-frequency parasitic oscillations (which usually lie in the wide range between 100 and 2000 kc.) invariably involve plate- and grid-circuit r.f. chokes in combination with a split-stator tank condenser tuning at least one of them if not both. The normal tank coils have such little reactance at low frequencies that they may be considered merely as long connecting leads.

Although they are not so likely to be encountered in amplifiers using the better-screened transmitting tetrodes and pentodes, low-frequency parasitic oscillations are often found in stages employing triodes and the less effectively-shielded audio tubes, such as the 6L6, 6V6, etc. Even if well-screened tubes are used, it is safer and more convenient to arrange the circuit in advance so that these low-frequency circuits are broken up.

Circuits To Be Avoided

Fig. 6-33 shows several commonly-used circuit arrangements that should be avoided to eliminate the possibility of low-frequency parasitics. In A, either r.f. choke or both may be replaced with a 100-ohm resistor, as shown in B. In a similar circuit, parallel feed can be used in either grid or plate, but not in both.

In Fig. 6-33C, RFC_2 should be replaced by a resistor. If parallel plate feed is desired, series feed should be used in the grid, as shown at D, necessitating parallel feed in the driver-stage plate. If the driver plate tank circuit has a split-stator condenser, as shown in E, the grid choke should be replaced by a 100-ohm resistor by-passed to ground, as shown in F. It is important that the by-pass be fairly large so as to be effective at low frequencies.

A check for low-frequency parasitics should be made after the v.h.f. oscillations have been eliminated. The check is conducted along the lines described for very-high frequencies. Low-frequency oscillation can be detected by coupling the absorption wavemeter closely to the r.f. chokes involved, remembering that the range of frequencies over which this type of parasitic may occur is wide. They can also sometimes be detected by listening on a receiver close to the transmitter, when harmonics, usually rough in character, may be heard at regular intervals that are multiples of the fundamental frequency. On a calibrated receiver, the fundamental frequency can be determined by observing the spacing between adjacent harmonics.

Neutralizing Procedure

The procedure in neutralizing is essentially the same for all types of tubes and circuits. The filament of the amplifier tube (or tubes) should be lighted and excitation from the preceding stage fed to the grid circuit. There should be no plate voltage on the amplifier.

The immediate objective of the neutralizing process is reducing to a minimum the r.f. driver voltage fed from the input of the amplifier to its output circuit through the grid-plate capacitance of the tube. This is done by adjusting the neutralizing condenser until an r.f. indicator in the output circuit gives minimum response.

NEUTRALIZING INDICATORS

Fig. 6-34 shows the diagram of a sensitive neutralizing indicator. By referring to Chapter Sixteen, it will be seen that this forms part of the indicating absorption wavemeter also recommended for checking parasitic oscillation. The link should be coupled to the output tank coil at the low-potential or "ground" point. Care should be taken to make sure that the coupling is loose enough at all times to prevent burning out the meter or the rectifier.

A neon bulb touched to the "hot" end of the tank coil will glow if enough feed-through voltage is developed across the tank, but it is a less-sensitive device. Another disadvantage is that its use introduces capacitance across one side of the circuit which may unbalance the circuit, thus giving an inaccurate indication of neutralization.

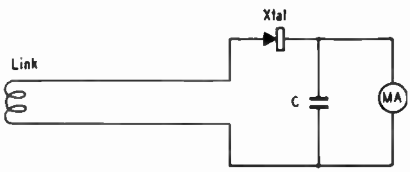


Fig. 6-34 — Circuit of sensitive neutralizing indicator. *Xtal* is a 1N34 crystal detector, *MA* a 0-1 direct-current milliammeter and *C* a 0.001-μfd. mica by-pass condenser.

A more satisfactory indicator than the neon bulb is a flashlight bulb (the lower the power the more sensitive) connected at the center of a turn or two of wire coupled to the tank coil at the low-potential point. Its sensitivity is poor compared with the milliammeter-rectifier indicator, however.

The grid-current milliammeter may also be used as a neutralizing indicator. If the amplifier is not neutralized, there will be a large dip in grid current as the plate-tank tuning passes through resonance. This dip reduces as neutralization is approached until at exact neutralization all change in grid current should disappear.

NEUTRALIZING ADJUSTMENTS

The neutralizing condenser should always be adjusted with an insulating rod, not only to protect the operator but also to avoid capacitive effects which might give a false indication.

With excitation applied, the neutralizing adjustment should be started with the neutralizing condenser at minimum capacitance, increasing the capacitance in small steps. At each step, the plate tank should be swung through resonance which will be indicated by maximum deflection of the indicators mentioned above and by the dip in grid current. As the point of

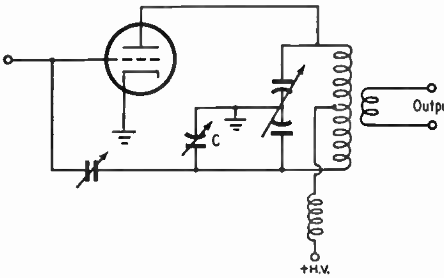


Fig. 6-35 — In this neutralizing circuit, *C*, which should have the same capacitance as the output capacitance of the tube, has been added to compensate for the tube capacitance across the upper half of the circuit.

neutralization is approached, the indication will become less until it is a minimum when neutralization is reached. If the neutralizing capacitance is increased further, the indication will again increase. If the neutralizing condenser has a proper range of capacitance, it should always be possible to find a point of minimum indication with an increase on either side.

If it is found that neutralization does not hold over the entire range of the tank condenser for any one band in a single- or parallel-tube amplifier, the balancing condenser of Fig. 6-35 should be added and adjusted.

In an amplifier which is to be used on several bands, it should be first neutralized when

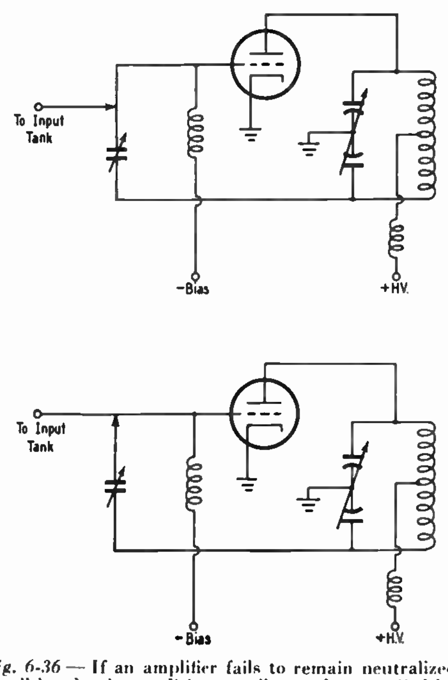


Fig. 6-36 — If an amplifier fails to remain neutralized on all bands, the condition usually can be remedied by tapping the input-tank lead along the neutralizing-condenser lead (or vice versa), adjusting the position until the amplifier neutralizes at the highest frequency at the same setting of the neutralizing condenser as at the lowest frequency. The same adjustment should be made to both sides of a push-pull circuit.

tuned to the lowest-frequency band. Then the neutralization should be checked at the highest frequency. If it is found that the neutralizing condenser needs readjustment at the higher frequency, the connection between the grid tank circuit (or the plate tank circuit of the driver with capacitance coupling) should be adjusted as indicated in Fig. 6-36 until the neutralizing adjustment is the same for both bands, always neutralizing first on the lowest-frequency band and checking at the highest frequency. If there are parasitic chokes at the grid and plate, connection of the neutralizing condenser to one side and then the other should be tried to determine which connection permits the best neutralizing from band to band.

Any indication remaining at minimum means that coupling between input and output exists external to the tube. The isolation seldom can be made complete, but it should be possible to bring it down to a very low value with proper wiring and shielding. Short leads in neutralizing circuits are highly desirable, and the input and output inductances should be so placed with respect to each other that magnetic coupling is minimized. Usually this requires that the axes of the coils be at right angles to each other. In some cases it may be necessary to shield the input and output circuits from each other. Magnetic coupling can be detected by disconnecting the plate tank from the remainder of the circuit and testing for r.f. in it as the tank condenser is tuned through resonance. The driver stage must be operating while this is done, of course.

Adjustment of Inductive Neutralizing Systems

With link neutralizing of a single-tube or parallel-tube amplifier, the neutralization can be adjusted by altering the number of turns at either end or by changing the spacing between the link and the tank coil.

The inductive neutralizing system holds neutralization over a wider frequency range if the auxiliary adjusting condenser (Fig. 6-17C) is omitted, adjusting the size of the coil to resonate at the operating frequency with the plate-grid capacitance of the tube only. The use of the auxiliary condenser makes the adjustment more convenient, of course.

Harmonic Reduction

A transmitter may generate and radiate energy at harmonics of the operating frequency of any stage. Although this harmonic power seldom is large in terms of the power at the output frequency, nevertheless it may be sufficient to cause interference at considerable distances under favorable conditions of propagation, as well as to various amateur and nonamateur services, particularly television, at shorter distances.

Low-frequency harmonics seldom are of great consequence, provided that the tank circuits in the transmitter have a reasonable amount of capacitance, that link coupling is used at both input and output of the final amplifier, and that the antenna system is provided with a tuned circuit. V.h.f. harmonics that fall in television channels, however, are a matter of considerable concern to a constantly-increasing number of amateurs, principally because the distance involved between transmitter and television receiver may be so small.

The primary means of reducing TVI, so far as the transmitter proper is concerned, may be divided into three general parts. First, the transmitter circuits and the manner of operating the tubes may be adjusted to bring the amplitude of harmonics generated in the transmitter to a minimum. Second, the transmitter may be shielded to prevent direct radiation of harmonic energy from the transmitter components and wiring. The third step that may be taken is that of filtering the power leads entering the shielding enclosure to prevent radiation by the external power leads. One or all of these steps may be necessary. The extent to which treatment must be carried will vary according to the proximity of the receiver and the strength of the television signal. The last two steps mentioned must be considered as one in practice, since one is seldom effective unless accompanied by the other.

Circuit and Layout

At the outset, the use of link coupling between all stages of a transmitter is strongly recommended wherever harmonics must be reduced to the extreme minimum. In any case, it should be used between the driver and the final amplifier. Link lines should be made with coaxial cable.

The transmitter should be designed so that the frequency-multiplier stages can be operated at low power level without the need for driving any stage excessively hard. The required step-up in power should be secured through the use of straight amplifiers at the operating frequency. In c.w. transmitters, the output stage can be operated with cut-off bias, or even somewhat less, without a drastic reduction in efficiency. When the output stage is to be plate-modulated, the driving power and bias should be reduced to the minimum that will give linear operation at the plate input used.

In laying out the components of a transmitter for minimum harmonics, the same precautions should be observed as those recommended in avoiding v.h.f. parasitic oscillation in a preceding section of this chapter. The arrangement should be such that all r.f. leads are as short as possible. This will help to move the resonant frequency of lead segments and combinations with shunting capacitances up above the television channels. Where a lead cannot be shortened physically, its inductance can be reduced by the use of large conductor, such as copper strip or tubing. By-pass condenser leads should be as short as possible and the condensers should be grounded to the chassis at points as close as possible to the cathode or filament grounding point. Mica condensers should be used, since their inductance is less than that of conventional paper condensers.

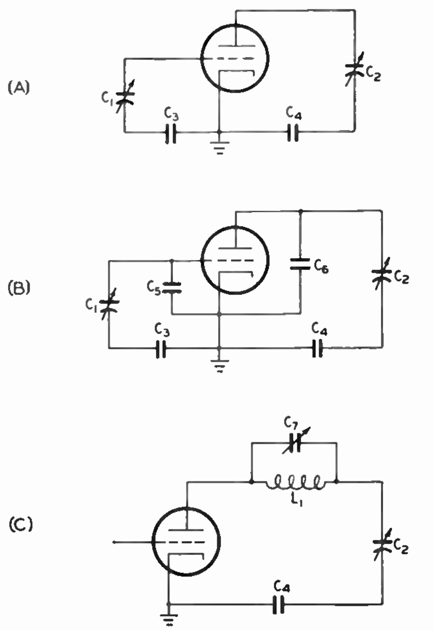


Fig. 6-37 — A — A common circuit for harmonic resonance. B — Tube-shunting condensers added to shift resonances. C — Plate-circuit wavetrap used to break up or shift v.h.f. resonance. C_1 and C_2 are the usual input and output tank condensers, C_3 and C_4 the respective by-pass condensers. C_5 may be a 10- to 15-μμfd. ceramic condenser wired directly across the tube socket. C_6 is a tubular air condenser or one of the high-voltage vacuum type with heavy leads. A capacitance of 10 to 15 μμfd. is usually satisfactory here also. C_7 and L_1 usually are made to tune through the second and third harmonics of 28-Mc. band frequencies. C_7 can be a 100-μμfd. midget and L_1 may consist of 3 turns No. 14 wire, ½-inch diameter, ½ inch long.

After a stage has been wired up, it should be checked for resonances in the television-frequency range with a grid-dip meter. If resonances are found, they can often be shifted to a higher or lower frequency by a change in the length of leads in the offending circuit.

One of the most troublesome circuits is the path through the tank condenser (either grid or plate or both) as shown in Fig. 6-37A. When these paths are found to be resonant in the television band, the resonant frequency often can be changed beneficially by the addition of fixed shunting condensers connected directly from plate or grid to cathode or filament. Since these shunting condensers are in parallel with the tank condensers, the resonant frequency of the combination will be lower than that of the original resonance, while the frequency of the new path through the fixed condensers will be higher, because the inductance involved is small. To achieve this, the inductance of the shunting condensers should be as small as possible. In the grid circuit, small ceramic condensers of 10 to 15 μμfd. have been found satisfactory when the driving power does not exceed that obtainable from one or two 807s. In the plate circuit tubular air or high-voltage vacuum condensers connected with heavy leads should be used. These condensers should have a peak-voltage rating of twice the d.c. plate voltage for c.w. operation, or of four times the d.c. plate voltage with plate modulation.

In other cases it may be necessary to break up the resonance, or shift its frequency, with a wavetrap, as shown in Fig. 6-37C. Although sometimes required for satisfactory reduction in harmonics in the transmitter tank, this step is not a highly desirable one, however, since it will have an influence on the neutralizing adjustment in triode amplifiers and on the balance in push-pull amplifiers using either triodes or screen-grid tubes. Furthermore, the selectivity of the trap is sufficient to make it necessary to retune the trap for any appreciable change in operating frequency.

Since even mica condensers possess some inductance, they should be checked for resonances after they have been wired in the circuit. In some cases, it may be beneficial to shunt a mica by-pass condenser with a low-capacitance ceramic unit, or even a tubular air condenser if the voltage requires it.

All r.f. connections, especially those to the chassis, should be solid, since a loose connection may cause rectification and consequent generation of harmonics. Where parts of the r.f. circuit (the tube base, for instance) are below the chassis, while the remainder is above, all connections should be made to both top and bottom surfaces of the chassis.

Power-Lead Filtering

Filtering of power leads from the transmitter is of utmost importance, since any harmonic energy permitted to flow back through the power lines is easily conducted or radiated to near-by television receivers. The use of shielded wire for all power wiring inside the transmitter has been found to be very effective in reducing harmonics in external power leads. Such wire not only is shielded against pick-up of r.f., but it also may act to attenuate harmonics through its continuous capacitance to ground.

The type of filter that experience has thus far shown to be most effective for high-voltage and bias leads is shown in Fig. 6-38A. It consists of a v.h.f. choke and a mica condenser on the terminal side, or on both sides. In some cases the filtering may be found better with the condenser on the terminal side only.

Except possibly in extreme cases, it is sufficient to filter the a.c. line to the filament-transformer primary if the transformer is so placed that the secondary leads are not too long. The filtering arrangement is shown in Fig. 6-38B. Condensers of the "feed-through" type (such as the Sprague "Hypass") have a definite advantage over ordinary mica condensers in this application. The chokes will not always be found necessary.

All power-lead filters should be located as close to the point of exit from the chassis as possible and the components should be shielded from any possible r.f. pick-up. If there are no r.f. circuits under the chassis, the shielding offered by the chassis itself may be sufficient. But if r.f. circuits are placed under the chassis, the filters should be covered with separate shields. It is also important that the leads external to the chassis be shielded until they leave the transmitter enclosure. Unshielded wires should not be exposed to direct r.f. fields as shown in Fig. 6-39. If a filament center-tap or cathode lead is brought out, it too should

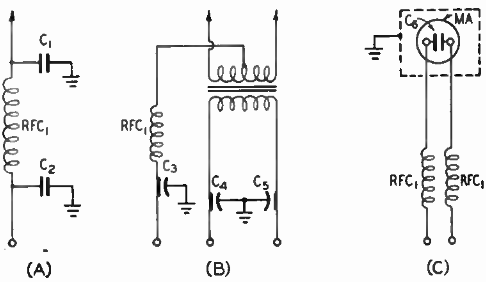


Fig. 6-38 — V.h.f. filters in power leads. A — Filter for plate- or bias-supply lead. B — Filament-circuit filtering. C — Meter filtering. C_1, C_2 and C_6 should be 470-μμfd. mica condensers. Feed-through type condensers (Sprague "Hypass") are recommended for C_3, C_4 and C_5. Capacitances of 0.01 to 0.1 μfd. have been found satisfactory. RFC_1 is a 7-μh. v.h.f. choke (Ohmite Z-50).

be filtered, as shown in Fig. 6-38B, with a v.h.f. choke and a "feed-through" type condenser.

Meters should be enclosed in shielding covers and connected with shielded wire. A mica by-pass condenser should be strapped across the meter terminals and v.h.f. chokes placed as shown in Fig. 6-38C at the point where the meter connections to the circuit are made. Any meter multipliers must, of course, be adjusted with the meter wired up and in place.

Care should be used in the selection of shielded wire for transmitter use. Not only should the insulation be conservatively rated for the d.c. voltage in use, but the insulation should be of material that will not easily deteriorate in soldering. For high voltages, automobile ignition cable covered with shielding braid is recommended. The shield of all wiring should be grounded as often as a convenient grounding point can be found. Where the

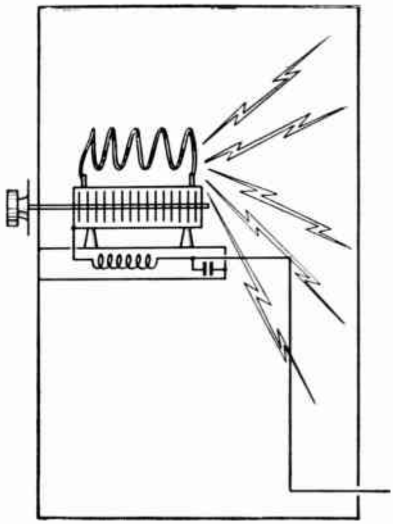


Fig. 6-39 — A metal cabinet can be an adequate shield, but there will still be radiation if the leads inside can pick up r.f. from the transmitting circuits.

power wiring crosses or runs parallel, the shields should be spot-soldered together.

Transmitter Enclosure

The transmitter itself should be operated in a metal enclosure. This of course means that the panels must be of metal. While commercial cabinets and cabinet-type racks do not provide "tight" shielding, the screening usually will be sufficient if the other measures outlined are taken. Homemade enclosures of copper screening, as described at the end of this chapter, can be made to provide adequate shielding and ventilation at the same time.

With the transmitter operating and coupled to a dummy load as shown in Fig. 6-40, harmonic fields in proximity to the transmitter enclosure and energy flowing in the power leads can be checked with a sensitive absorption wavemeter of the type described in Chapter Sixteen. The wavemeter should be tuned

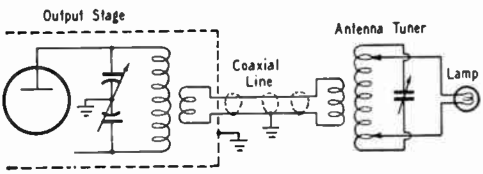


Fig. 6-40 — Loading arrangement which should be used in checking harmonics around transmitter and in power leads.

to the harmonic frequencies that may fall in the TV bands. By means of these checks, it will be possible to determine the effect of the measures taken. When checking power leads external to the transmitter enclosure, a small section of the power lead should be formed into a loop which should be coupled tightly to the coil of the wavemeter.

A Simple Single-Tube Transmitter

One of the simplest practical transmitters is shown in the photographs of Figs. 6-41 and 6-42. If the station receiver has a power audio stage which is not required for headphone reception, the tube may be taken from the receiver and used in the transmitter (provided that the tube is a pentode or tetrode as it usually is). A plug inserted in the empty socket in the receiver may be used to obtain power for operating the transmitter.

The schematic diagram is shown in Fig. 6-43. The Tri-tet oscillator circuit is used to permit operation in either the 3.5- or 7-Mc. bands with a single 3.5-Mc. crystal. Series plate feed is used and no means of reducing the voltage of the screen below that of the plate is necessary if the supply potential does not exceed 250 to 300 volts.

The cathode circuit is tuned by a fixed mica condenser, C_1, but if necessary, the tuning of this circuit can be changed by changing the dimensions of the coil, L_1.

No provision is included for tuning the antenna system, for the sake of maximum simplicity. This can be done by selecting the proper feeder length and adjusting the size of the antenna coupling coil, L_3.

Construction

To minimize the tools required for the construction of the transmitter the parts are mounted on a simple chassis of wood finished with clear lacquer or shellac. Two 1¾ × 9¾-inch strips of ¼-inch-thick wood are fastened with screws to the two 4½ × 2½ × ¾-inch end pieces, leaving enough separation between the strips for the Amphenol M1P octal sockets used for holding both the crystal and the tube. Wood screws can be used to mount the sockets, or they can be bolted to the wood strips with 6-32 machine screws. The key of the tube socket should be mounted toward the front of the transmitter for convenience in wiring the plate circuit to the tuning condenser. Because the tuning condenser does not have a long mounting shank, it is necessary to drill a clearance hole for the shank and then dig away — or counterbore — clearance for the nut. The two Fahnestock clips for the antenna are secured under two of the screws used for fastening the wood strips to the right-hand end piece, and the other two clips used for the key leads are held down by machine screws on the left-hand end piece. The r.f. choke is held in place on the left-hand end piece by a machine screw. The four wires used for a power cable are brought out at the rear left under the wood strip — a half-round hole is filed in the end piece to clear the wires.

The plate and antenna coils are held in place on three small sticks set in the top of the chassis — penny suckers are a good source of these sticks. The bottom of the plate coil connects to a brass machine screw soldered to a lug which is sweated to the stator terminal of the tuning condenser, and the screw is built up most of its length by adding nuts or small spacers. The screen end of the coil, the top end of the winding, is fastened to a brass screw that runs through the rear wood strip. The coil ends have lugs soldered to them to facilitate band changing. The antenna-coil ends similarly fasten to two brass screws supported by short lengths of heavy wire and the wire is sweated to the Fahnestock clips and to the heads of the screws.

Wiring

The wiring is done with the same wire that is used for the coils, because a single 50-foot

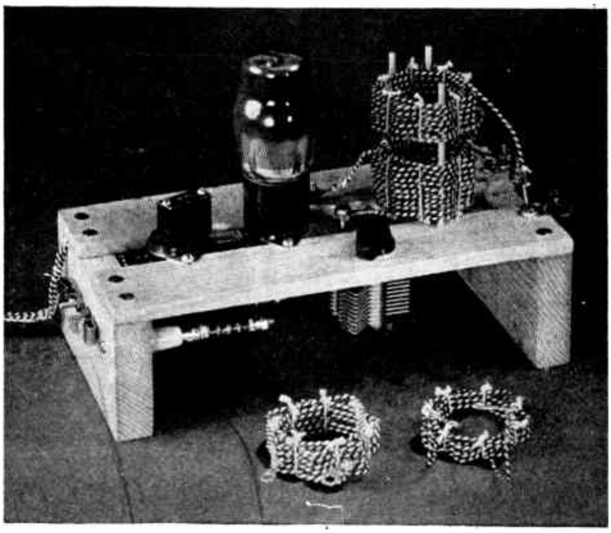


Fig. 6-41 — By using wood for the chassis and simplified construction throughout, this simple oscillator transmitter can be built with very few shop tools. Using a 3.5-Mc. crystal, operation in the 3.5- and 7-Mc. bands is possible by changing the plate and antenna coils. The arrangement is suitable for 6F6, 6V6 or other similar pentodes and tetrodes.

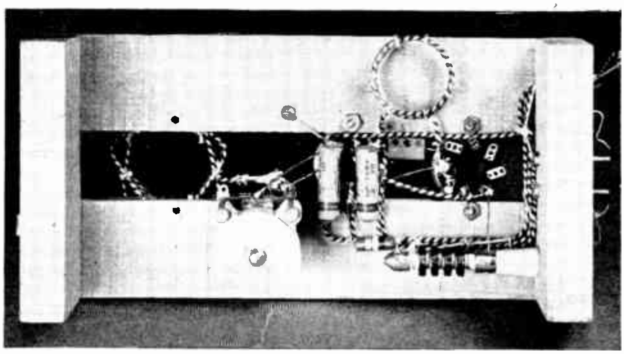


Fig. 6-42 — Bottom view of the simple single-tube transmitter. The cathode coil is between the tube and crystal sockets. The r.f. choke is to the right, C_4 is at left center with the two by-pass condensers, C_2 and C_3, to the right of it.

roll of No. 18 bell wire, available in any "5 & 10" or hardware store, suffices for the whole rig with some to spare. To insure good electrical connection, the wire is soldered at every connection, which means that the wire is soldered to the heads of the brass machine screws used for the key leads and the screen end of L_2 before the screws are put in place. One key lead, one end of R_1, the outer foil connections on C_2 and C_3, and the lead to Pin 1 of the power plug must be connected to Pin 1 of the tube socket. At the crystal socket, two adjacent pins (e.g., 1 and 8) are bonded together for the grid side of the crystal and the next two pins (e.g., 2 and 3) are bonded together for the cathode side. This permits plugging the crystal into either Pins 8 and 2 or 1 and 3. The connection can be elaborated still further by bonding Pins 4 and 5 with 8 and 1 and tying 6 and 7 to 2 and 3, in which case the crystal can be plugged in any way and it will make the proper connection.

The cathode coil, consisting of 5 turns of No. 18 bell wire, is wound on a 1¼-inch diameter form and then removed and tied with string at a number of places. The cathode coil is mounted by its leads only but, being short, they offer adequate support.

The plate and antenna coils are wound by equally spacing seven nails on a 2-inch diameter circle, driving the nails completely through the board used so that the heads are flush against the board. Small spikes can be used, or nails of the "8-penny" size will be satisfactory if a thin board is used. One end of the wire is secured to a nail and the wire is threaded over alternate nails, so that the coil repeats itself every two turns. When the required number of turns has been made, the end of the wire is wrapped around a nail and the coil tied together with string at the seven cross-over points. Soldering lugs are soldered to the ends of the coil for ease in changing bands.

The four wires coming out the side of the chassis that go to the power plug are twisted together slightly and cabled with string to form a neat cable, and the cable plug, P_1, is simply the base from an old tube. If the receiver is to be used as a source of power, the base should be one that will fit the power-output tube in the receiver. Break the tube and chew out the glass from inside the base with a pair of pliers, being careful not to break the bakelite of the base. It will help in making connection to the proper pins if a small drill, slightly larger than the diameter of the No. 18 wire, is run through the pins before the wires are inserted and soldered in place.

Tuning

After checking the wiring, plug in a crystal and connect the 7-Mc. coil in place. Place the audio tube from the receiver in the transmitter and plug in the power cable, and connect a key to the clips on the side of the transmitter. If the receiver has push-pull output, it is probably best to remove both power tubes. Set the tuning condenser, C_4, at about 40 per cent meshed and turn on the power to the receiver. When the tube has had time to warm up — about 30 seconds — close the key and touch a neon bulb to the plate end of L_2. Or a small 10-watt electric lamp can be connected to the antenna posts with the 6-turn antenna coil in place. If C_4 is set properly, the

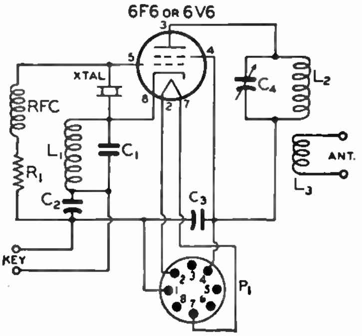


Fig. 6-43 — Wiring diagram of the inexpensive easy-to-build transmitter.

C_1 — 470-μμfd. mica.
C_2, C_3 — 0.01-μfd. 600-volt paper.
C_4 — 140-μμfd. variable (Hammarlund SM-140 or Bud MC-1876).
R_1 — 0.1-megohm 1-watt composition.
L_1 — 5 turns No. 18 d.c.c., 1¼-inch inside diameter, close-wound.
L_2 — 3.5 Mc.: 19 turns. 7 Mc.: 12 turns.
L_3 — 13 turns and 6 turns. Requires experiment — see text. See text for L_2 and L_3 winding instructions.
P_1 — See text.
RFC — 2.5-mh. r.f. choke (National R-100U).

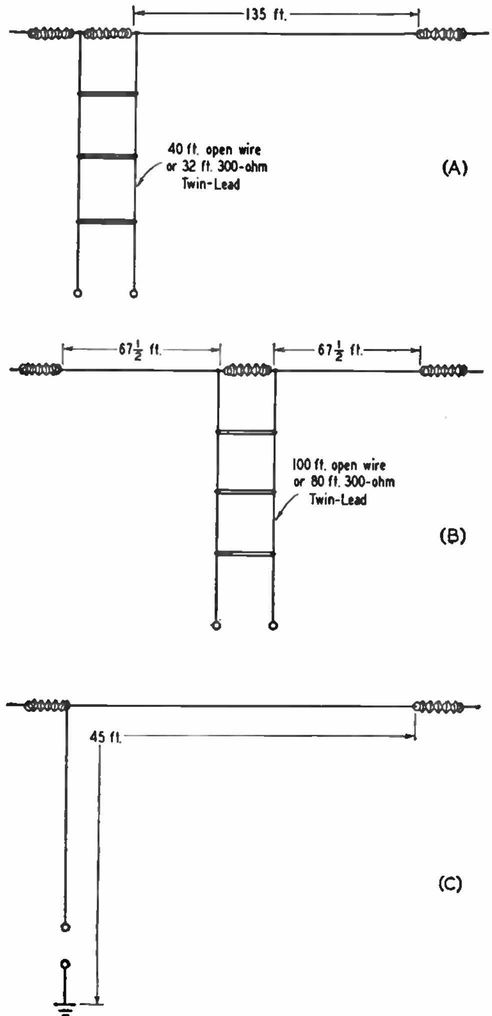


Fig. 6-44 — **Suggested antenna dimensions for use with the single-tube transmitter. A — End-fed half-wave or Zepp. B — Center-fed half-wave. C — Quarter-wave grounded antenna.**

neon bulb will glow or the lamp will light. If this does not happen, try tuning the plate condenser until signs of output become apparent. The transmitter can then be checked on the 3.5-Mc. band by putting in the proper coils — remembering, however, to turn off the receiver and hold the key closed until the power pack of the receiver has been discharged, to avoid getting a shock when touching the coil terminals. The tuning condenser setting will be about 85 per cent meshed on the lower-frequency band.

It will not be possible in most cases to check the keying on the receiver used to furnish power to the transmitter, and it is highly advisable to check the keying in a monitor or another receiver. If the keying is chirpy, the cathode coil, L_1, should be squeezed out of round to reduce its inductance until the keying is better. On the 3.5-Mc. band, best keying will generally be obtained with slightly less capacity at C_4 than the setting for maximum output. In the oscillator shown in the photographs, a slight key click on "break" was eliminated by connecting a 0.1-μfd. 600-volt paper condenser directly across the key. Some crystals key better than others.

Antennas

A 135-foot piece of wire for the antenna can be fed in several ways to give satisfactory results. It can be fed at one end with about 40 feet of open-wire feeders (about 32 feet of Amphenol 300-ohm Twin-Lead), as shown in Fig. 6-44A or it can be fed in the center with 100 feet of open-wire feedline (about 80 feet of 300-ohm Twin-Lead) as indicated at B. These lengths will enable one to connect the feedline directly to the antenna posts of the transmitter without the necessity for tuning condensers — other lengths may require either series or parallel condensers. Some experiment with the antenna coil may be necessary, but a small flashlight bulb in series with one of the feeders will serve as a good indication of feeder current, and will help in the tune-up process. The lamp need not be shorted during normal operation unless it burns too brightly. A neon bulb will also help in detecting r.f. energy in the transmission line, but it may not always light with this low power.

If room for only a short length of wire is available for the antenna, say 40 or 50 feet, it is best to connect its end to one antenna post and a good ground to the other as shown in Fig. 6-44C. Here again some experimentation will be necessary to determine the optimum size of L_3. The diagram of a suitable alternative power supply is shown in Fig. 6-45.

The power can be increased by substituting a 6L6 for a smaller tube and adding a separate power supply to give 350 volts at 100 ma., but with the newer small crystals it is not advisable to increase the voltage much above this value without keeping the screen voltage down by the addition of a dropping resistor and another by-pass condenser.

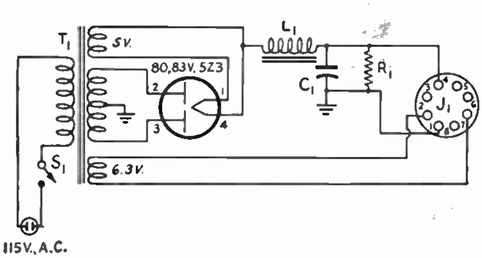


Fig. 6-45 — **Circuit diagram of alternative power supply for the simple single-tube transmitter.**

C_1 — 8-μfd. 450-volt electrolytic.
R_1 — 25,000 ohms, 10 watts.
L_1 — Filter choke — any receiver replacement type, 15 hy. or more, 50 ma. or more.
J_1 — 8-prong tube socket.
S_1 — S.p.s.t. toggle switch.
T_1 — Power transformer — any receiver replacement type, not over 750 volts c.t., 50 ma. or more.

A Low-Power VFO Transmitter for the 3.5- and 7-Mc. Bands

A complete 20-watt c.w. transmitter for the 80- and 40-meter bands is shown in Figs. 6-46 through 6-52. Considerable economy and simplification in both circuit and operation result by confining the function to these two bands. Additional considerations are that the need for v.h.f. filtering and other measures against TVI is eliminated for most localities and that the special design often found necessary for satisfactory VFO performance when operating at the higher frequencies is not required. The transmitter has been designed to make maximum use of the capabilities of any available low-cost receiver-type power supply delivering from 150 or 200 volts to 350 or 400 volts. Since both bands are covered with a single buffer-doubler coil, only two plug-in coils are required. Instability at the operating frequency, often experienced with the less-expensive receiver audio screen-grid tubes, is eliminated because the input of the buffer-doubler is untuned and the output stage always operates as a push-push frequency doubler. This type of stage is capable of efficiency approaching that of a straight amplifier, without the usual high driving-power requirements usually associated

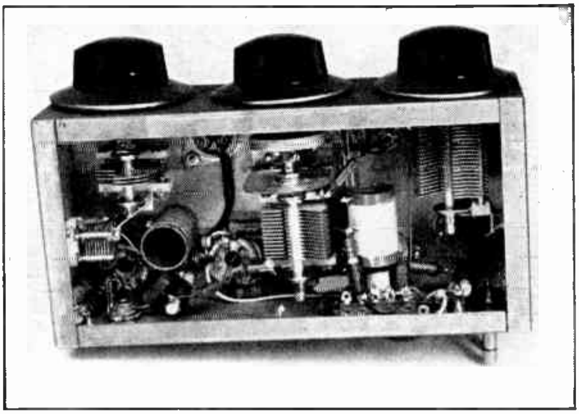


Fig. 6-47 — Bottom view of the r.f. section of the low-power transmitter for 3.5 and 7 Mc. The oscillator is to the left and the output tank condenser to the right. The oscillator and buffer-doubler coils should be spaced away from the chassis and mounted at right angles.

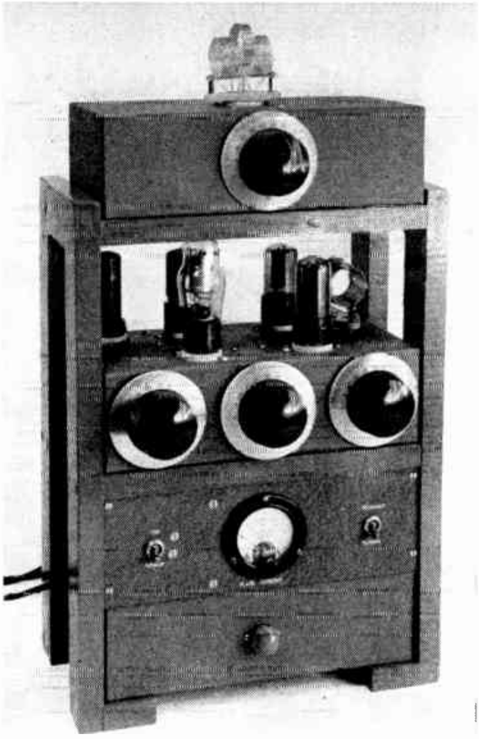


Fig. 6-46 — The four units of the low-power transmitter are assembled in a simple frame of 1 by 2 wood strips. The rack measures 16 inches high, 11¾ inches wide and 6¼ inches deep.

with a conventional single-tube doubler. Higher-order harmonic output also is comparable with that of a straight-through amplifier stage.

Circuit Details

A 6AG7 is used in the series-tuned Colpitts VFO circuit and 6V6s are used in both buffer-doubler and output stages. Type 6F6s or 6L6s may be used instead of 6V6s without changes in circuit values. C_1 is the main oscillator tuning control, while C_2 serves to set the tuning range over the proper frequencies. The oscillator delivers output at 1.75 Mc. at all times. The buffer-doubler stage is tuned to 1.75 Mc. for 3.5-Mc. output and to 3.5 Mc. for 7-Mc. output from the transmitter.

Parallel plate feed is used in the output stage to remove high voltage from the plug-in coil. The oscillator and output stages are keyed simultaneously in the common cathode lead, while the buffer-doubler stage is protected with cathode-resistor bias. A switch on the control panel opens up the cathode circuit of the output stage while the VFO is being set to frequency. A VR-150 is included on the chassis to regulate the screen voltage for the oscillator and buffer-doubler and the plate voltage for the 6AG7.

Construction

The complete transmitter is made up of four separate units assembled on a simple framework or rack made of 1 by 2 wood strips. The units from bottom to top are power supply,

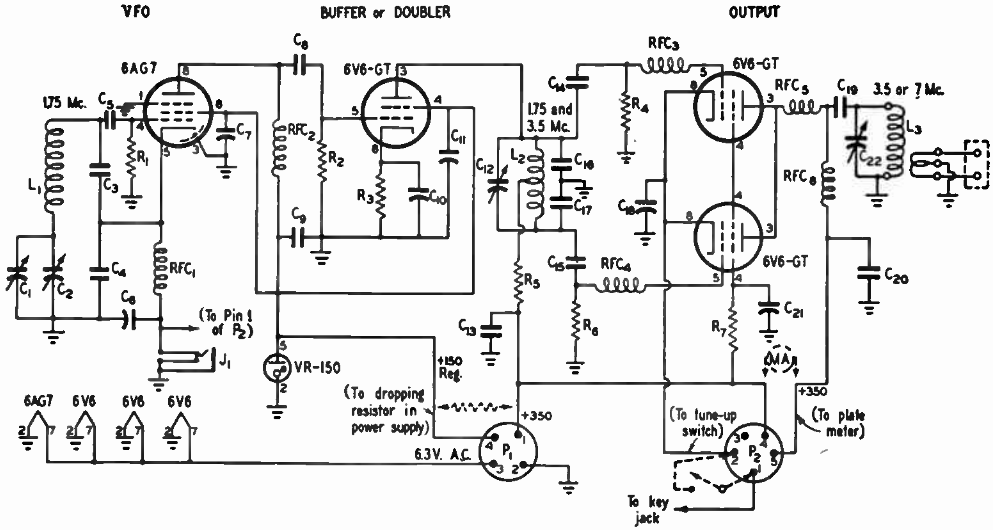


Fig. 6-48 — Circuit of the low-power c.w. transmitter for the 3.5- and 7-Mc. bands.

C_1 — (Osc. tuning) approx. 40-$\mu\mu$fd. variable (Millen 19050 with one rotor plate removed).
C_2 — (Bandset) 50-$\mu\mu$fd. max. midget variable (National PSR-50).
C_3, C_4 — 1000-$\mu\mu$fd. silver mica.
C_5, C_8, C_{14}, C_{15} — 100-$\mu\mu$fd. mica.
C_6, C_7, C_{10}, C_{11}, C_{18}, C_{21} — 0.01-μfd. paper, 600 volts.
C_9, C_{13}, C_{19}, C_{20} — 1000-$\mu\mu$fd. mica.
C_{12} — 200-$\mu\mu$fd. variable (Millen 19200).
C_{16}, C_{17} — 22-$\mu\mu$fd. mica.
C_{22} — 325-$\mu\mu$fd. variable (Millen 19325).
R_1, R_2, R_4, R_6 — 47,000 ohms, ½ watt.
R_3 — 470 ohms, 1 watt.
R_5 — 150 ohms, 1 watt.
R_7 — 10,000 ohms, 5 watts.
L_1 — 95 turns No. 32 d.s.c. close-wound on 1-inch diam. form.
L_2 — 48 turns No. 24 d.s.c. close-wound on 1-inch diam. form.
L_3 — 3.5 Mc.: 14 μh. (National AR-17-40E). 29 turns No. 20, 1¼-inch diam., spaced to occupy 1½ inches. 4-turn center-tapped link.
— 7 Mc.: 4 μh. (National AR-17-20E). 14 turns No. 18, 1¼-inch diam., spaced to occupy 1¼-inch length. 4-turn center-tapped link.
J_1 — Closed-circuit 'phone jack.
P_1 — 4-prong male connector.
P_2 — 5-prong male connector.
RFC_1, RFC_2, RFC_6 — 2.5-mh. r.f. choke (National R-100-S).
RFC_3, RFC_4 — 1-μh. r.f. choke (National R-33).
RFC_5 — 16 turns No. 20 d.s.c. close-wound on ¼-inch diam. form. (A 1-watt resistor of any high value may be used as the form.)

control-and-meter panel, r.f. section, and antenna coupler. The r.f. section, as well as the power supply and antenna tuner, is built on a 5 × 10 × 3-inch chassis. The three tuning condensers are mounted with their shafts in line, equally spaced along the front of the chassis. The rotor of C_{12} is insulated from the chassis by mounting the condenser on a small subpanel of polystyrene and fitting the shaft with an insulating coupling. The bandset condenser, C_2, is mounted on the left-hand end of the chassis so that it may be adjusted with a screwdriver from outside. All tube sockets and the socket for the plug-in output tank coil are submounted. L_1 and L_2 are wound on Millen 1-inch diameter forms and are fastened permanently under the chassis with their axes at right angles. Two plugs on cords are provided — one with four pins for the power-supply connection and one with five pins for the control and meter connections. A vernier dial (National type AM) is used for the oscillator. Matching straight dials (National type P) are used for the other tuning controls.

The diagram of the power supply is shown in Fig. 6-52. With condenser input, this particular unit will deliver between 350 and 400 volts under load. R_1 is the voltage-dropping resistor for the VR tube. The transformer and rectifier tube are placed at one end of the chassis and the other components underneath

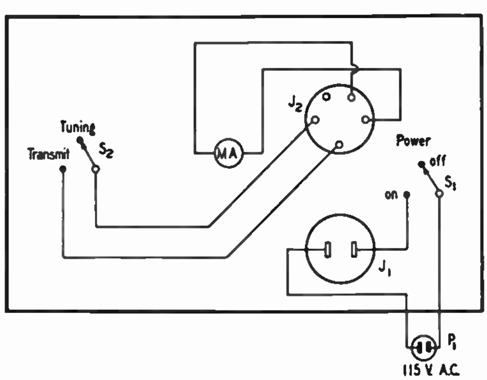


Fig. 6-49 — Wiring diagram of the meter and control panel for the 3.5- and 7-Mc. transmitter.

J_1 — Female a.c. receptacle.
J_2 — 5-prong female receptacle.
MA — 0-200 ma. d.c.
P_1 — Male a.c. plug.
S_1, S_2 — S.p.s.t. toggle switch.

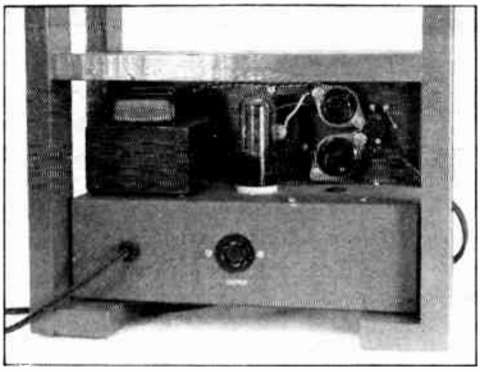


Fig. 6-50 — Rear view of the power supply and control panel in place in the frame.

so that the two sockets on the control panel may be reached from the rear. In mounting the transformer, room between the transformer and the front edge of the chassis should be allowed for the toggle switch at the right-hand end of the control panel. The latter is not fastened to the chassis but to small wood strips on the frame.

In addition to the two toggle switches and the meter at the front, two sockets are mounted at the back of the 3⅝ × 10-inch control panel, toward the left-hand end. The plug from the r.f. section goes in the five-prong socket and the a.c. cord from the power supply plugs into the a.c. outlet. The wiring of this panel is shown in Fig. 6-49.

The diagram of the antenna coupler is shown in Fig. 6-51. Connections between the coil and the base pins may be made appropriate for either series or parallel tuning, whichever may be found necessary, as shown in the circuit. The proper connections are then made automatically when the coil is plugged into the socket. The rotor of the antenna

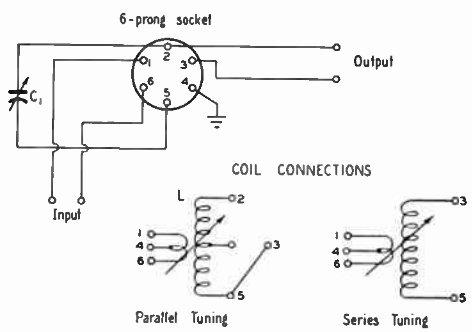


Fig. 6-51 — Wiring diagram of the antenna tuner for the 3.5- and 7-Mc. c.w. transmitter. The coil connections can be altered as shown to provide either series or parallel tuning when the coil is plugged in.

C_1 — 200-$\mu\mu$fd. variable (Millen 19200).
L — 3.5 Mc. — 22 turns No. 22, center-tapped, 1¼-inch diam., 1⅝ inches long, 6-turn center-tapped link (National AR-17-40S).
— 7 Mc.: 12 turns No. 18 center-tapped, 1¼-inch diam., 1½ inches long, 4-turn center-tapped link (National AR-17-20S).

tuning condenser is insulated from the chassis in the manner described for the buffer-doubler tank condenser.

Adjustment

With the amplifier switch (S_2, Fig. 6-49) off, the VR voltage-dropping resistor (R_1, Fig. 6-52) should be adjusted so that the VR tube just ignites with the key closed. Then the oscillator bandset condenser, C_2, should be adjusted while listening on a receiver so that the signal is heard at 3500 kc. with C_1 set near maximum capacitance. With the amplifier switch turned on, and the 80-meter coil in place, C_{12} should be adjusted to the point near maximum capacitance where the milliammeter shows a slight increase in current. Then the output-stage tank condenser, C_{22}, should be adjusted for minimum plate current to the

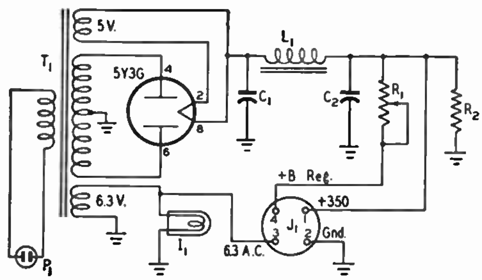


Fig. 6-52 — Circuit diagram of the power supply for the low-power transmitter.

C_1, C_2 — 8-μfd. 450-volt electrolytic.
R_1 — 10,000 ohms, 50 watts, with slider.
R_2 — 50,000 ohms, 10 watts.
L_1 — 8 hy., 100 ma.
I_1 — 6.3-volt pilot-lamp assembly.
J_1 — 4-terminal female connector.
P_1 — Male a.c. connector.
T_1 — 375-0-375 v., 100 ma.; 5 v., 3 a.; 6.3 v., 4 a. (UTC R-9).

final (approximately 10 ma. unloaded). The adjustment for 7-Mc. output is similar, except that the buffer-doubler tank condenser will be set near minimum capacitance and the 7-Mc. coil will be plugged into the output stage. If, by any chance, it is found that both bands are not covered, it will be necessary to adjust the size of L_2 by a couple of turns, keeping the same number of turns on each side of the center-tap. In adjusting the transmitter, care should be used not to tune up on the third harmonic of the oscillator at 5.2 Mc. This resonance will be found near the center of the range of C_{12}.

With c.w. operation, it is permissible to load the output stage until it draws 100 ma., even though the power transformer may have a rating of 100 ma. or somewhat less. A plate voltage of 400 should not be exceeded, but the transmitter will operate satisfactorily (at reduced input, of course) from power supplies having much lower voltage and current ratings.

For adjustment of the antenna tuning and coupling, see "Practical Coupling Systems," Chapter Ten.

A Two-Control 75-Watt All-Band Plug-In-Coil Transmitter or Exciter

Through the use of bandpass couplers and fixed-tuned circuits, the number of tuning controls necessary for the adjustment of the transmitter shown in Figs. 6-53 through 6-57 is reduced to two — one for setting the frequency of the VFO and the other for resonating the final output tank circuit. The general system of operation is as follows: A 6AG7 series-tuned Colpitts VFO operating at 1.75 Mc. doubles frequency in the plate circuit, driving a second 6AG7 as a buffer at 3.5 Mc. or as a doubler to 7 Mc. The output of the latter stage can be connected (at W) to drive the 807 output stage for either 3.5- or 7-Mc. output (W connected to Y), or to drive a third 6AG7 (W connected to X). Through the use of plug-in coils, the third 6AG7 may be operated as a doubler to 14 Mc., a tripler to 21 Mc., or a quadrupler to 28 Mc. to drive the output stage on any one of these bands (W connected to X, and Z connected to Y for 14 Mc.; W connected to X and L_7 connected to Y for 28 Mc.). When the plug Z is not in use, it is plugged into a jack-top insulator mounted nearby so that its capacitance to ground will be held constant. If this precaution is not taken, its position may affect the tuning.

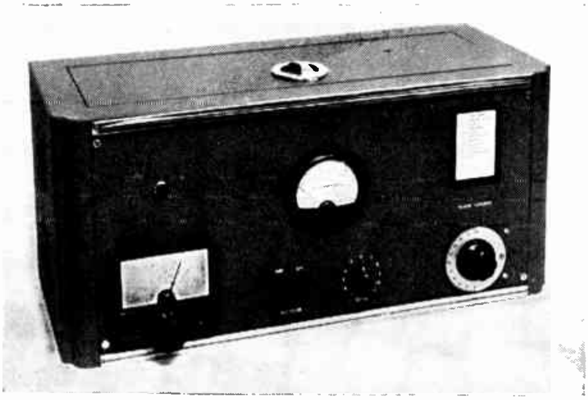


Fig. 6-53 — The VFO tuning dial and the amplifier tuning-condenser control are at the left and right ends of the panel, respectively. Switches for the metering circuit and for control of the multiplier heater circuit are at the center of the panel just below the meter.

Circuit Details

A bandpass coupler, L_2C_3-L_3C_4, is used between the oscillator and the first 6AG7. This coupler cuts off rather sharply on either side of a substantially-flat range from 3.37 to 4 Mc. to minimize undesired oscillator and harmonic drive to the buffer stage.

The plug-in coils (L_4) for the plate circuit of the second stage are approximately self-resonant. Four of these are required — two for driving the 807 at 3.5 and 7 Mc., one for driving the following frequency multiplier for optimum output over the 14-, 21- and 28-Mc. bands, and the fourth for use when final-stage output is desired over the full range of the 27-Mc. band.

Self-resonant coils (L_5) are also used in the output of the third 6AG7 multiplier for 14, 21 and 27 Mc., but a bandpass coupler (C_5L_6-C_6L_7) is used for 28 Mc. both to obtain the desired bandwidth and to reduce v.h.f. harmonics. This stage, when not in use, is disabled by turning off the heater by throwing switch S_2.

Standard plug-in coils cover all bands in the 807 output stage, a single coil serving for the 21-, 27- and 28-Mc. bands. RFC_4 is a parasitic suppressor. C_{30} is a tubular fixed air condenser connected directly from plate to cathode with short leads. This condenser not only aids in suppressing parasitic oscillation but also helps to reduce harmonic output. C_7 and L_8 make

Fig. 6-54 — Rear view of the 75-watt all-band transmitter or exciter. The v.h.f. trap is fastened to the top of the tubular condenser, C_{30}, to the rear of the output tank coil and alongside the 807 and the 6V6GT. A small baffle shield separates the two multiplier coil sockets, shown here with the 28-Mc. bandpass coupler in place. The mounting of the low-frequency coupler is shown between the oscillator tube and the two frequency multipliers.

up a v.h.f. harmonic wavetrap that may be tuned to either the second or third harmonic of 28-Mc. band frequencies. All power leads are completely filtered for protection against radiation of high-frequency harmonics.

The 6V6GT provides automatic protection for the 807 when excitation is removed. A VR tube also is included on the chassis to regulate the common plate and screen voltage for the oscillator. S_1 is a meter switch for checking the plate, screen and grid currents of each stage. The switch leads also are filtered for high-order harmonics.

Construction

The construction departs somewhat from the conventional in that the components are assembled on an aluminum bottom plate for the 7 × 17 × 3-inch chassis, the inverted chassis serving as a bottom cover. The aluminum is easier to work than steel.

The rear view shows the oscillator tube at the right-hand end of the chassis with the low-frequency bandpass assembly to the left. A slot, 1¼ by 2½ inches, is cut in the aluminum plate to allow clearance for the filter compo-

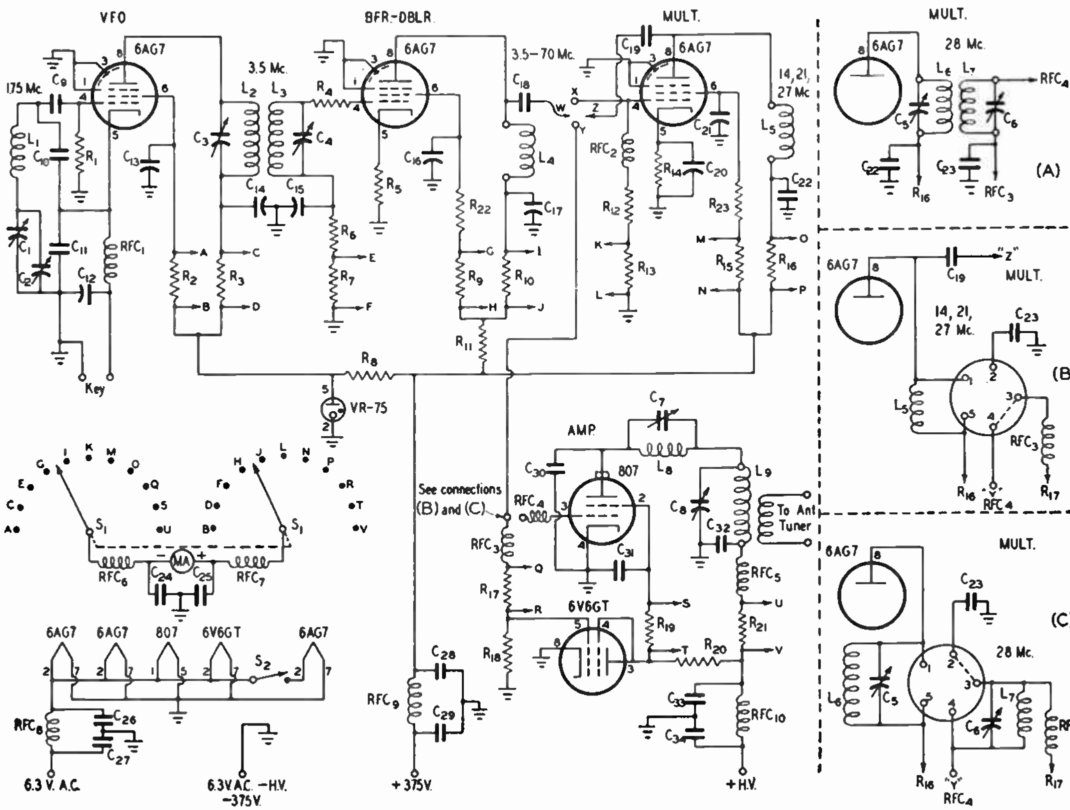


Fig. 6-55 — Circuit diagram of the two-control all-band transmitter.

C_1 — 50-μμfd. variable (Millen 19050).
C_2 — 100-μμfd. variable (Millen 20100).
C_3, C_4, C_5, C_6 — 5–20 μμfd. ceramic trimmer (Centralab 820B).
C_7 — 100-μμfd. air trimmer (Millen 26100).
C_8 — 250-μμfd. variable (National TMS-250).
C_9, C_{18} — 100-μμfd. mica.
C_{10}, C_{11} — 680-μμfd. silver mica.
C_{12}, C_{13}, C_{14}, C_{15}, C_{16}, C_{17}, C_{20}, C_{21} — 0.01-μfd. paper, 400 volts.
C_{19} — 15-μμfd. mica.
C_{22}, C_{31} — 0.001-μfd. mica.
C_{23} — 680-μμfd. mica.
C_{24}, C_{25}, C_{26}, C_{27}, C_{28}, C_{29} — 470-μμfd. mica.
C_{30} — 12 μμfd. (Millen 15015).
C_{32} — 0.01-μfd. mica, 1200 volts.
C_{33}, C_{34} — 340-μμfd. mica (two 680-μμfd. units in series).
R_1 — 47,000 ohms, ½ watt.
R_2, R_3, R_7, R_9, R_{10}, R_{13}, R_{15}, R_{16}, R_{17}, R_{19} — 100 ohms, ½ watt.
R_4 — 47 ohms, ½ watt.
R_5, R_{14} — 330 ohms, 1 watt.
R_6, R_{18} — 22,000 ohms, ½ watt.
R_8, R_{11} — 10,000 ohms, 10 watts.
R_{12} — 0.1 megohm, ½ watt.
R_{20} — 75,000 ohms, 20 watts (two 10-watt resistors in series).
R_{21} — Meter shunt: 51 inches No. 28 wound on a high-resistance ½-watt resistor.
R_{22} — 33,000 ohms, 1 watt.
R_{23} — 16,500 ohms (two 33,000-ohm 1-watt in parallel).
L_1 to L_9, inc. — See coil table.
MA — 0–50 d.c. milliammeter.
RFC_1, RFC_2, RFC_3, RFC_5 — 2.5-mh. r.f. choke.
RFC_4 — 1-μh. r.f. choke (National R33).
RFC_6, RFC_7, RFC_9, RFC_{10} — 7-μh. r.f. choke (Ohmite Z-50).
RFC_8 — 36 turns No. 18 enam., 9/16-inch diam., close-wound on National PRE-3 form.
S_1 — 2-pole 2-section 11-position selector switch (Centralab 1413).
S_2 — S.p.s.t. rotary toggle switch.

COIL TABLE FOR TWO-CONTROL ALL-BAND TRANSMITTER

Coil	*L in µh.*	*Wire*	*Turns*	*Diam., In.*	*Length, In.*	*Coml. Type*
L_1	92	30 s.s.c.	68	1	Close-wound	On Millen 45000 form
L_2, L_3	57	30 enam.	44	1	Close-wound	On Millen 45000 form
L_4 (A)	51	28 enam.	50	1	Close-wound	On Millen 45004 form
(B)	14.3	22 enam.	26	1*	Close-wound	On Millen 45004 form
(C)	13.6	22 enam.	25	1*	Close-wound	On Millen 45004 form
(D)	16	22 enam.	28	1	Close-wound	On Millen 45004 form
L_5 (E)	5.3	22 enam.	13	1	Close-wound	On Millen 45005 form
(F)	2.45	22 enam.	9	1	7/16	On Millen 45005 form
(G)	1.1	22 enam.	6	1	3/8	On Millen 45005 form
L_6 } (H)	1.27	22 tinned	6	1	3/8	B&W 3015 Miniductor
L_7 }	0.76	22 tinned	4	1	1/4	B&W 3015 Miniductor
L_8		16 enam.	4	5/16	3/4	
L_9 (J)	11	16 tinned	22 (8)	1 1/2	2	Millen 43042
(K)	6.1	16 tinned	13 (8)	1 1/2	1 1/4	Millen 43042 with 9 turns removed
(L)	2.55	14 tinned	9 (2)	1 1/2	1 1/2	Millen 43022
(M)		14 tinned	4 (2)	1 1/2	1 3/8	Millen 43012

* End turns adjustable — see text.
NOTE: Figures in parentheses after turns for L_9 are link turns. Links wound over L_9 at ground end. Adjust as necessary.

COIL LINE-UP FOR TWO-CONTROL TRANSMITTER

Band	L_4	L_5	L_6L_7	L_9
3.5	A	—	—	J
7	C	—	—	K
14	B	E	—	L
21	B	F	—	M
27	D	G	—	M
28	B	—	H	M

nents. The buffer-doubler tube, the frequency-multiplier tube and the VR-75 form a line from rear to front just to the left of the filter. The oscillator bandset condenser, C_2, is to the right of the VR-75. Coils for the driver tubes and a crystal socket (contacts X and Y in the circuit diagram) are next in line to the left. The 807 amplifier tube, the tubular by-pass capacitor, C_{30}, the harmonic trap, and the tank coil are at the left end of the chassis. Connections to points W and Z in the circuit diagram go through feed-through bushings between the 807 and the driver-coil sockets. The stand-off insulator mounted in front of the 807 is used as a low-capacity holder for terminal Z when inductive coupling from the multiplier tube is used. The antenna terminals are to the rear of the tank coil and the power-cable terminals to the rear of the oscillator tube.

All power wiring should be done with shielded wire.

Bandpass Couplers and Coils

The windings of the self-resonant coils should not be cemented in place until they have been finally adjusted in the circuit by spreading or compressing the turns. Also, means should be provided for adjusting the coupling between the two coils in each of the bandpass couplers. In the case of the low-frequency coupler, which is wound on a 1-inch form, one of the windings can be wound over a layer of paper between the wire and the form so that it may be slid back and forth. The coil

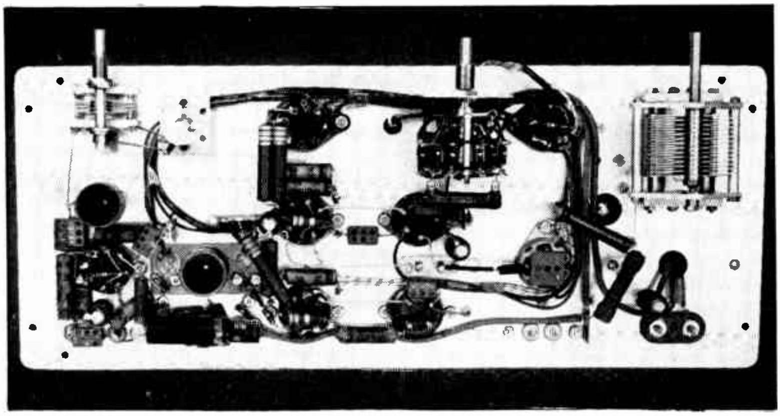


Fig. 6-56 — Bottom view of the all-band plug-in-coil transmitter. The oscillator coil, tuning condenser and bandset condenser are in the upper left-hand corner, with the low-frequency bandpass coupler below and to the right. The middle shaft is that of the meter switch mounted on a metal bracket. The output tank condenser is to the right.

Fig. 6-57 — Close-up view of the 28-Mc. bandpass coupler.

form and the two trimmer condensers are fastened to a 2 × 2½-inch piece of sheet polystyrene. A rectangular hole cut in the chassis permits the unit to be mounted with the condensers on top, where they may be adjusted, and the coil at the center underneath.

The 28-Mc. coupler components also are mounted on a piece of polystyrene, this one measuring 1½ by 2¾ inches. This is then fastened to a small stand-off insulator mounted on a 5-pin plug so that the unit may be plugged into the 5-prong coil socket. After the coupling has been adjusted, the coils may be cemented to the base. Fig. 6-55A shows the circuit of the last 6AG7 when the coupler is being used for 28-Mc. output, while B and C show the connections to the coil sockets for either a self-resonant coil (B) or the coupler (C). Dotted lines indicate pins that should be wired together in the *coil plug* rather than in the socket.

Adjustment

The unit is designed to operate from two power supplies. The one for the oscillator and multiplier stages should deliver 375 volts at 100 ma. or more. The 807 may be operated from any supply delivering from 450 volts or so to a maximum of 750 volts, the output obtainable being in proportion. A 6.3-volt filament transformer, rated at 3 amperes or more, is required for the heaters.

The oscillator tuning range should first be set, by adjusting C_2, so that it covers 1.68 to 2 Mc. with C_1. The low-frequency coupler can be adjusted by observing the grid current to the 807 with the 3.5-Mc. coil plugged in at L_4. With the oscillator set at 3.37 Mc., C_4 should be adjusted for maximum grid current. Then, with the oscillator set at 1 Mc., C_3 should be adjusted for maximum grid current. If, on checking, as the oscillator is tuned across the band the grid current shows a pronounced peak at one end or the other, the size of L_4 in the buffer stage should be adjusted slightly to bring resonance farther away from the end of the band at which the peak occurs. If there is a decided dip in grid current between the two ends of the band, the coupling between the two coupler coils may be too great or too little. With the circuits tuned as described and the coupling adjusted to the proper point, the grid current to the 807 stage should be essentially flat over the desired band and drop off rather rapidly at each end. Once the coupler has been adjusted for the 3.5-Mc. band no further adjustment should be required for the higher frequencies, the adjustment for the latter being taken care of by adjusting the self-resonant plate coils to maintain grid current to the 807 as constant as possible over the band in each case. The 28-Mc. coupler is adjusted in the same manner described for the low-frequency unit. Under all conditions, the grid current to the 807 amplifier should average 3 to 4 ma. For proper adjustment of the 807 output circuit, see "Adjustment of R.F. Amplifiers," this chapter.

Current and Voltage Data

The plate and screen circuits of the oscillator should each draw approximately 3 ma. when the supply voltage is held at 75 volts by a regulator tube. The grid current for the next two 6AG7s should average 1 ma. Screen and plate currents of the buffer-doubler tube should be about 4 and 10 ma., respectively, and the screen and plate voltage should measure approximately 110 and 220 volts. Operating conditions for the screen of the frequency-multiplier tube are 7 ma. at 230 volts and the plate should draw about 20 ma. These figures can be expected to vary as the operating frequency of the transmitter is varied, because the self-resonant plate circuits will perform most efficiently over only a small band of frequencies. However, the meter readings should remain within a few per cent of the typical values listed above.

The screen of the 807 amplifier tube draws 5 to 6 ma. with an applied potential of approximately 300 volts. Normal full-load plate current for the 807 is 100 ma. and, with excitation removed, the 6V6GT should hold the d.c. input to less than 15 watts.

A 500-Watt Link-Coupled All-Band Transmitter

In the design of the transmitter shown in Figs. 6-58 through 6-69, an attempt has been made to incorporate means by which harmonic radiation and transmission may be minimized. In addition to the use of thorough shielding and power-line filtering, link coupling is used throughout.

Through the use of plug-in coils, the transmitter may be operated up to 21 Mc. with 1.75-Mc. crystals, and to 28 Mc. with either 3.5- or 7-Mc. crystals. With VFO input, it will go to 7 Mc. with 1.75-Mc. VFO output, to 21 Mc. with 3.5-Mc. VFO output, and to 28 Mc. with 7-Mc. VFO output.

The design of the push-pull triode final amplifier is suitable for any of the usual triodes with plate-cap connection, operating at plate voltages up to 1500 with plate modulation and a plate-voltage/total-plate-current ratio of 5 to 1 or greater.

The transmitter is made up in two sections mounted in a simple shielding enclosure consisting of a wood-strip frame covered with copper screening. The exciter unit is provided with pull handles and is designed to slide out for coil changing. As the unit is returned to the enclosure, the power-supply connections are automatically made at the rear through a series of plugs which fit into jacks set along the side of a 3 × 4 × 17-inch chassis fastened permanently at the rear. This chassis also encloses and shields the harmonic-filter components for all power-supply leads.

The second section above includes the push-pull final amplifier and an antenna tuner. The top cover is hinged to provide access to the output-stage and antenna tank coils. The meters for the amplifier stage are set in a separate panel between the two main sections.

Circuit Details

Referring first to the circuit diagram of the exciter section shown in Fig. 6-60, either the built-in Pierce crystal oscillator or an external VFO may be used to feed a 6L6 stage which is operated as a doubler, as a tripler, or, when necessary, as a buffer amplifier. This stage feeds a push-push 807 driver stage that may be operated either as a doubler, or as a self-neutralized straight-through amplifier by opening S_2 which controls the heater of one of the 807s. This inactive tube then becomes the

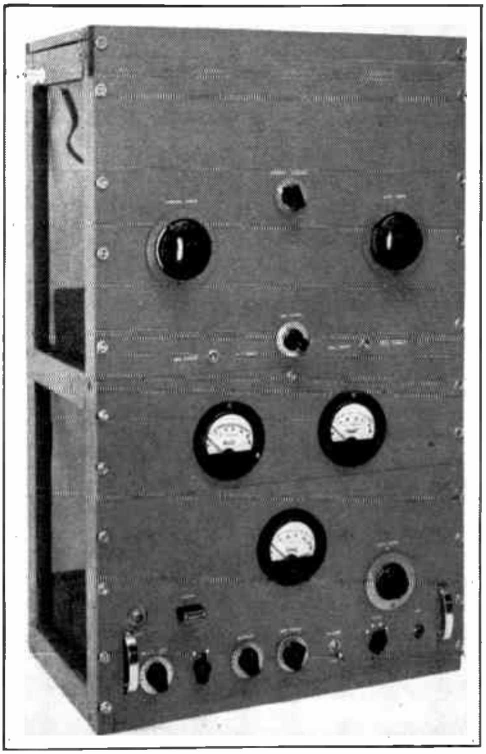


Fig. 6-58 — A complete 500-watt all-band transmitter including antenna tuner. The exciter unit at the bottom slides out for coil changing. The panel screws on this unit are dummies cemented in place. The top of the screened enclosure is hinged to permit changing coils in the final amplifier and antenna tuner.

Fig. 6-59 — Rear view of the completed 500-watt all-band transmitter with the back screening panel removed. The rectangular enclosed unit to the rear of the exciter contains the v.h.f. power-lead filters. The two matching boxes above enclose the amplifier-stage milliammeters.

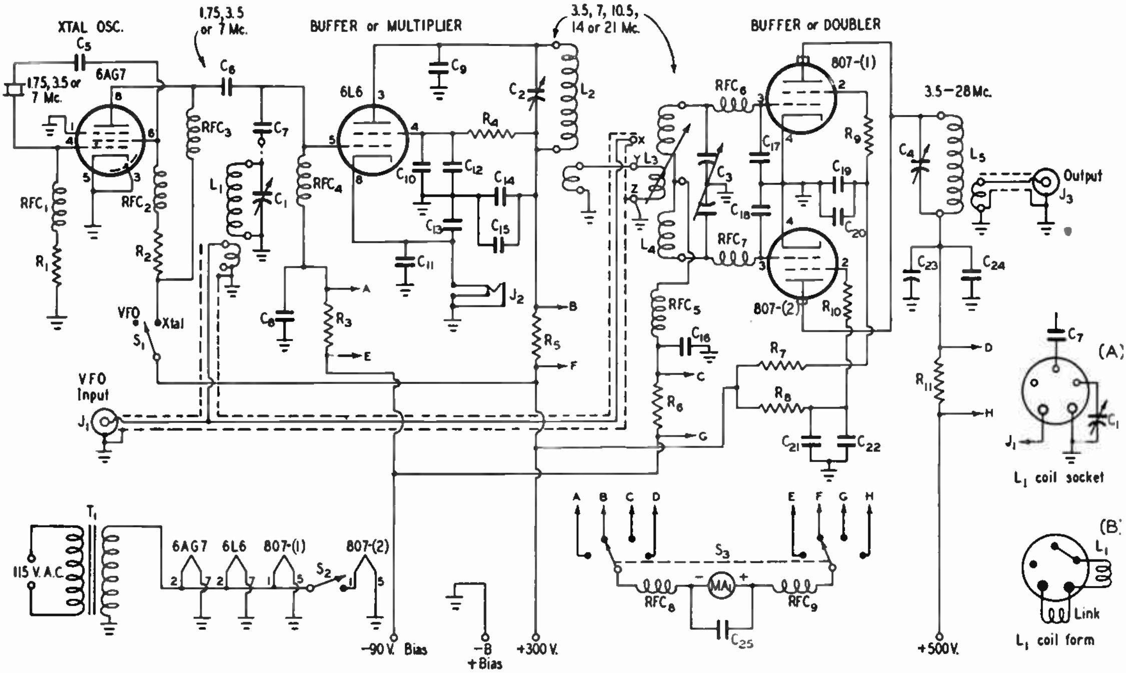


Fig. 6-60 — Circuit diagram of the exciter for the 500-watt all-band transmitter.

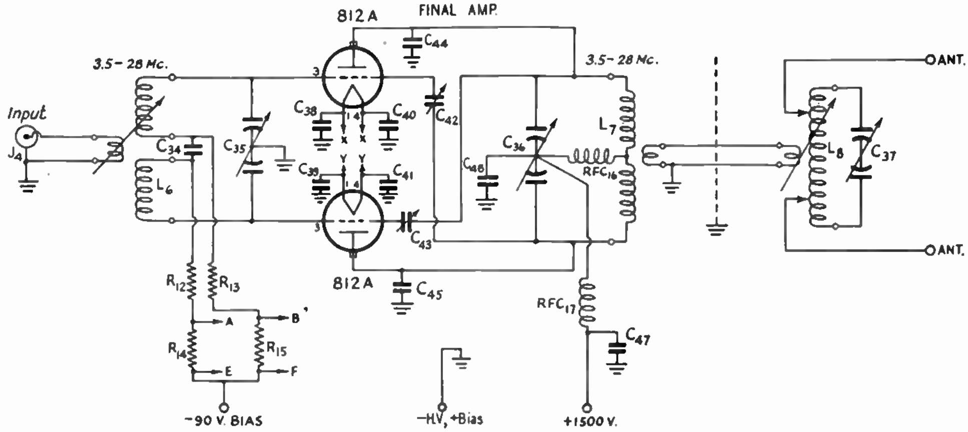
C_1, C_2 — 140-$\mu\mu$fd. variable condenser (Millen 22140).
C_3 — 100-$\mu\mu$fd.-per-section variable condenser (Millen 23100).
C_4 — 250-$\mu\mu$fd. variable condenser (National TMK-250).
C_5 — 0.0022-μfd. mica.
C_6, C_7 — 100-$\mu\mu$fd. mica.
C_8, C_{12}, C_{13}, C_{14}, C_{16}, C_{19}, C_{22} — 0.0047-μfd. mica.
C_9, C_{10}, C_{11}, C_{15}, C_{20}, C_{21} — 22-$\mu\mu$fd. ceramic.
C_{17}, C_{18} — 12-$\mu\mu$fd. ceramic.
C_{23} — 15-$\mu\mu$fd. air tubular (see text).
C_{24} — 0.001-μfd. 1200-volt-wkg. mica.
C_{25} — 470-$\mu\mu$fd. mica.
R_1 — 47,000 ohms, ½ watt.
R_2 — 5000 ohms, 2 watts.
R_3, R_6 — 100 ohms, ½ watt.
R_4 — 2500 ohms, 10 watts.
R_5, R_{11} — 10-times meter shunt (see text).
R_7, R_8 — 10,000 ohms, 10 watts.
R_9, R_{10} — 100 ohms, ½ watt, noninductive.
L_1, L_2, L_4, L_5 — See table.
L_3 — See line-up table for connections.
J_1, J_3 — Coaxial fitting.
J_2 — Closed-circuit jack.
MA_1 — Milliammeter — 25-ma. d.c. scale.
RFC_1 — 2.5-mh. 125-ma. r.f. choke.
RFC_2, RFC_3, RFC_4, RFC_5 — 2.5-mh. 50-ma. r.f. choke (National R-50).
RFC_6, RFC_7 — V.h.f. parasitic choke — 12 turns No. 16, ¼-inch diam., 1 inch long, self-supporting.
RFC_8, RFC_9 — 7-μh. r.f. choke (Ohmite Z-50).
S_1 — S.p.d.t. ceramic rotary.
S_2 — S.p.s.t. toggle.
S_3 — 2-pole 4-position 2-section ceramic rotary.
T_1 — Filament transformer: 6.3 volts, 6 amp.

COIL LINE-UP TABLE — 500-WATT ALL-BAND TRANSMITTER

Output	*XTAL*	*VFO*	L_1	L_2/ /L_4	L_3	L_5/ /L_6/ /L_7	S_2
3.5	1.75	—	1.75	3.5	Y-Z	3.5	Open
7	1.75	—	3.5	7	Y-Z	7	Open
14	1.75	—	3.5	7	Y-Z	14	Closed
21	1.75		3.5	10.5	Y-Z	21	Closed
3.5	3.5	—	None	3.5	Y-Z	3.5	Open
7	3.5	—	3.5	7	Y-Z	7	Open
14	3.5		3.5	7	Y-Z	14	Closed
21	3.5	—	3.5	10.5	Y-Z	21	Closed
28	3.5	—	7	14	Y-Z	28	Closed
7	7		None	7	Y-Z	7	Open
14	7	—	7	14	Y-Z	14	Open
28	7	—	7	14	Y-Z	28	Closed
3.5	—	1.75	1.75	3.5	Y-Z	3.5	Open
7	—	1.75	1.75	3.5	Y-Z	7	Closed
3.5	—	3.5	—	—	X-Z	3.5	Open
7	—	3.5	3.5	7	Y-Z	7	Open
14	—	3.5	3.5	7	Y-Z	14	Closed
21	—	3.5	3.5	10.5	Y-Z	21	Closed
7	—	7	—	—	X-Z	7	Open
14	—	7	7	14	Y-Z	14	Open
21	—	7	7	21	Y-Z	21	Open
28	—	7	7	14	Y-Z	28	Closed

"neutralizing condenser" for the other 807.

When the output frequency desired is the same as the crystal frequency, the input circuit of the 6L6 is not tuned, since this might result in instability in the 6L6 stage. In this case, no coil is used at L_1. When the coil socket and form of L_1 are wired as shown in Fig. 6-60A and B, the connection to tuning condenser C_1 is broken automatically when the coil is removed. When the VFO output frequency is the same as the desired frequency of operation, the VFO is fed directly to the input of the push-push stage through the link contacts at *X-Z* instead of *Y-Z*. The pins of the plug-in-coil base can be wired to make this connection automatic when the coil is plugged into place.



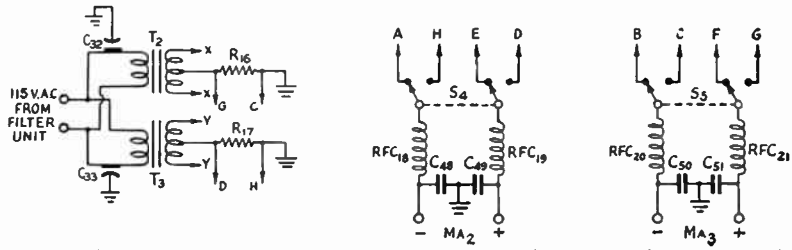


Fig. 6-61 — Circuit diagram of the final amplifier and antenna tuner of the 500-watt all-band rig.

C_{32}, C_{33} — 0.1 μfd., 250 volts (Sprague Hypass).
C_{34} — 0.0022-μfd. mica.
C_{35} — 100-μμfd.-per-section var. (Johnson 100HD-15).
C_{36}, C_{37} — 100-μμfd.-per-section variable (Johnson 100ED-30).
C_{38}, C_{39}, C_{40}, C_{41}, C_{48}, C_{49}, C_{50}, C_{51} — 47-μμfd. mica.
C_{42}, C_{43} — Neutralizing condenser — 1-11 μμfd. (Millen 15005).
C_{44}, C_{45} — 12-μμfd. 8000-volt tubular air condenser (see text).
C_{46}, C_{47} — 500-μμfd. 2500-volt-wkg. mica.
R_{12}, R_{13} — 1000 ohms, 10 watts (for 812A-s).
R_{14}, R_{15} — 10-times meter shunt (see text).
R_{16}, R_{17} — 100 ohms, ½ watt.
L_6, L_7, L_8 — See coil table.
J_4 — Coaxial connector.
MA_2, MA_3 — Milliammeter — 25-ma. d.c. scale.
RFC_{16} — 1-mh. 600-ma. r.f. choke (National R154).
RFC_{17}, RFC_{18}, RFC_{19}, RFC_{20}, RFC_{21} — 7-μh. r.f. choke (Ohmite Z-50).
S_4, S_5 — D.p.d.t. toggle switch.
T_2, T_3 — Filament transformer: 6.3 volts, 8 amp.

Double by-pass condensers are used at the cathode and screen terminals and for the plate-circuit returns in the 6L6 and 807 stages. A tubular air condenser is used at C_{23} to provide a low-inductance plate return to cathode. The larger by-pass condensers in each case are effective at the lower frequencies, but the ceramic condensers, together with RFC_6, RFC_7, R_9 and R_{10} in the 807 stage, are required to prevent v.h.f. parasitic oscillation. C_{10}, C_{17} and C_{18} are also used for the same purpose and to aid in the reduction of v.h.f. harmonics.

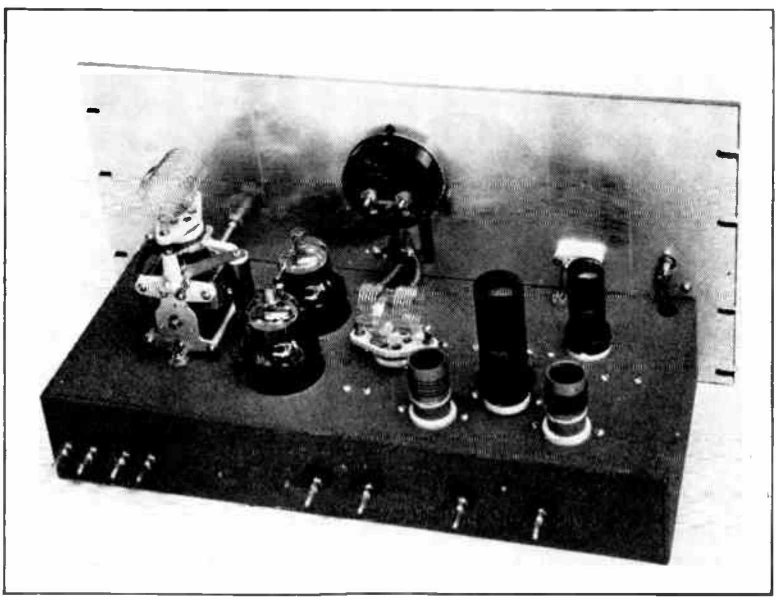


Fig. 6-62 — Rear view of the exciter of the 500-watt all-band transmitter.

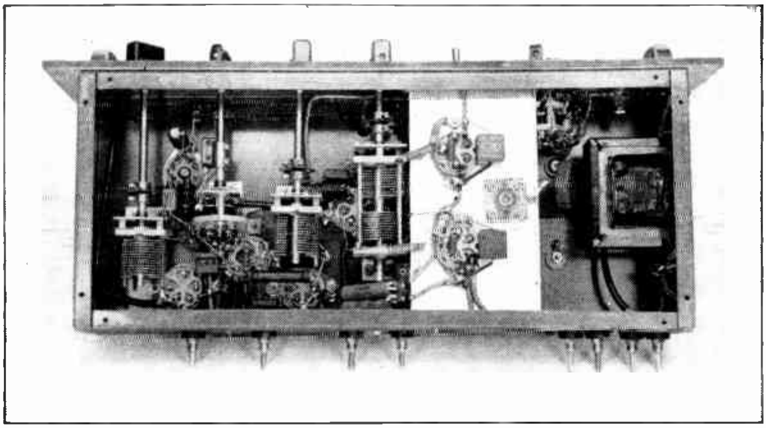


Fig. 6-63 — Bottom view of the exciter section of the 500-watt all-band transmitter.

which the two external milliammeters are connected by switches S_4 and S_5. RFC_{17} and C_{47} form one section of a v.h.f. filter in the high-voltage supply lead.

All other v.h.f. filters are in the separate shielded unit at the rear of the exciter. The wiring of this chassis is shown in Fig. 6-67.

R_3, R_5, R_6 and R_{11} are metering resistors across which the milliammeter may be switched by S_3 to read grid and plate currents in the two amplifier stages. R_5 and R_{11} should be adjusted in the circuit to give a scale multiplication of 10 times as described under "Measurement of Current, Voltage and Power," Chapter Sixteen. RFC_8, RFC_9 and C_{25} are part of the v.h.f. filtering system.

The circuit diagram of the push-pull final amplifier and the antenna tuner is shown in Fig. 6-61. The grid tank circuit is split for d.c. by the insertion of C_{34} at the center of L_6, and a separate filament transformer is used for each of the two tubes so that individual grid and cathode currents may be checked for amplifier balance. C_{44} and C_{45} are tubular air condensers connected directly with short leads from plate to ground near the tube sockets. They are essential in suppressing v.h.f. parasitic oscillation in this stage. R_{14}, R_{15}, R_{16} and R_{17} are metering resistors across

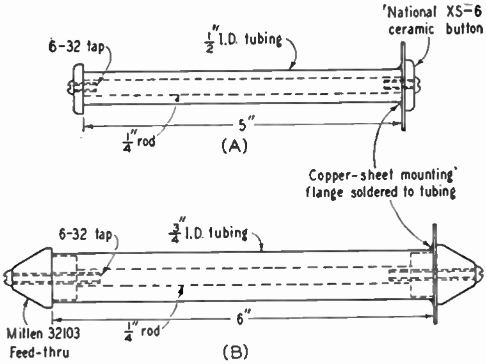


Fig. 6-65 — Sketches showing the construction of the air tubular condensers, (A) for the exciter and (B) for the final amplifier. The smaller condensers shown in the photographs of the final were replaced after preliminary tests.

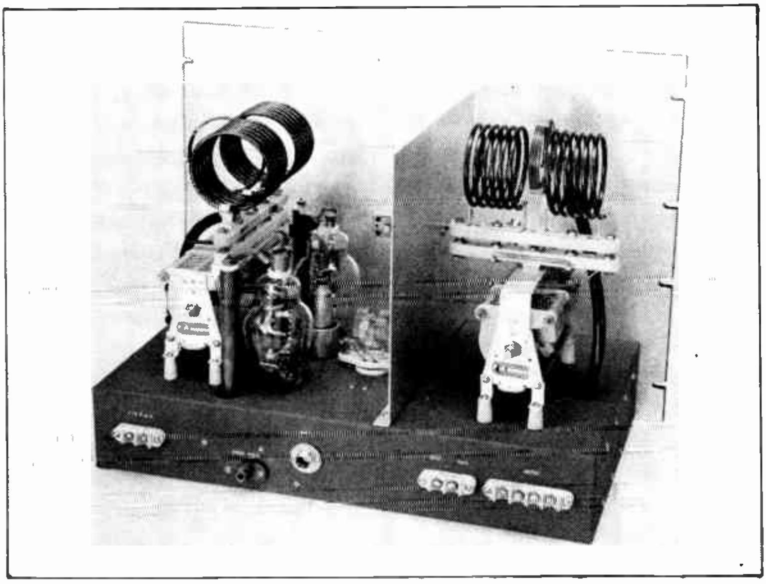


Fig. 6-64 — Rear view of the amplifier section of the 500-watt all-band transmitter. The antenna tuner is to the right. The gear box driving the variable link of the antenna tuner is fastened to the left-hand side of the shielding partition.

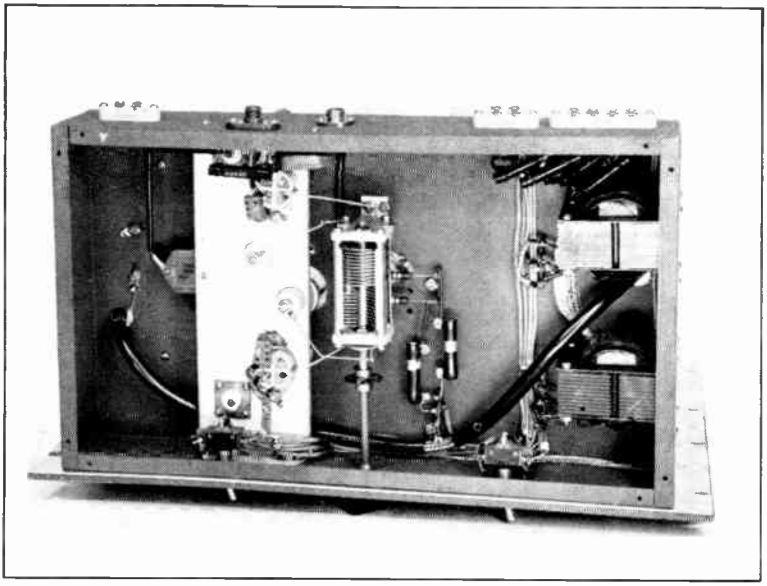


Fig. 6-66 — Bottom view of the amplifier section of the 500-watt all-band transmitter. The large coaxial lead runs from the amplifier link to the antenna-tuner link.

Exciter Construction

The exciter is assembled on a 7 × 17 × 3-inch chassis with a 10½-inch metal panel. The panel should be dropped so that its lower edge protrudes ¾ inch below the bottom edge of the chassis to cover the bottom strip of the frame of the enclosure. The arrangement of parts on top of the chassis is shown quite clearly in Fig. 6-62. The sockets for L_4 and L_5 are orientated so that the axes of the two coils are at right angles. Large clearance holes for the 807s and their shields, and one to clear the tubular condenser, C_{23}, also are cut in the top of the chassis. The 807 tank condenser, C_4, is insulated for d.c. by mounting it on polystyrene button-type insulators and providing an insulating coupling in the control shaft.

COIL TABLE — 500-WATT TRANSMITTER

Coil	*Band*	*Lμh.*	*Turns*	*Wire*	*Diam.*	*Lgth.*	*Link*	*Manufactured Type*
L_1/L_2	1.75	58	60	28 d.s.c.	1″	cw	8	Wound on Millen 45005 1-inch 5-pin bakelite form. See circuit diagram for pin connections.
	3.5	19	34	24 d.s.c.	1″	cw	6	
	7	7	18	22 d.s.c.	1″	cw	3	
L_2	10.5	4.4	16	22 d.s.c.	1″	⅞″	3	
	14	2.5	10	22 d.s.c.	1″	⅝″	3	
	21	1.2	7	18	1″	½″	2	
L_4	3.5	10	46	24	1¼″	1⅜″	10	National AR-17-80-S
	7	11	22	22	1¼″	1¼″	5	National AR-17-40-S
	10.5	8	18	22	1¼″	1″	5	National AR-17-40-S 2 turns off each end.
	14	2.9	12	18	1¼″	1⅛″	3	National AR-17-20-S
	21	1.3	6	18	1¼″	1⅛″	2	National AR-17-10S
L_5	3.5	10	22	16	1½″	1⅞″	3	B&W JEL-40
	7	3	12	14	1½″	2″	2	B&W JEL-20
	14	2.3	10	14	1½″	2¼″	2	B&W JEL-15
	21	0.8	6	14	1½″	2″	2	B&W JEL-10
	28	0.5	4	14	1½″	1¼″	2	B&W JEL-2 turns off
L_6	3.5	55	56	18	1¼″	1¾″	4	National AR-17-80C
	7	11	22	22	1¼″	1¼″	5	National AR-17-40S
	14	7	14	22	1¼″	⅞″	5	National AR-17-40S 4 turns off each side
	21	2.5	10	18	1¼″	1″	3	National AR-17-20S 1 turn off each side
	28	0.7	4	18	1¼″	½″	2	National AR-17-10S 1 turn off each side
L_7	3.5	40	40	14	2½″	5″	6	Johnson 500 HCF-80
	7	15	24	12	2½″	5″	6	Johnson 500 HCF-40
	14	3.7	12	6	2½″	5″	3	Johnson 500 HCF-20
	21	1	8	6	2″	5″	3	Johnson 500 HCF-10
	28	0.7	6	6	2″	4″	3	Johnson 500 HCF-10 1 turn off each side
L_8	Same as L_7, with swinging link							Johnson 500 HCS

Underneath, the tube sockets are mounted on a 3½-inch strip of aluminum spanning the bottom of the chassis. The tubular condenser also is fastened to this strip. The construction of this condenser is shown in Fig. 6-65A. In the push-push stage, the grid tank condenser, C_3, is immediately to the left of the tube sockets in Fig. 6-62 and at the center of the chassis, with the parasitic chokes, RFC_6 and RFC_7, fastened directly between the tube-socket and condenser terminals. Farther to the left in order are C_2, S_1, and C_1 with their shafts equally spaced.

To the right of C_3 are S_2, S_3 and the key jack, also equally spaced along the panel. The filament transformer is fastened to the right-hand end of the chassis. The crystal socket is mounted on the panel where it is

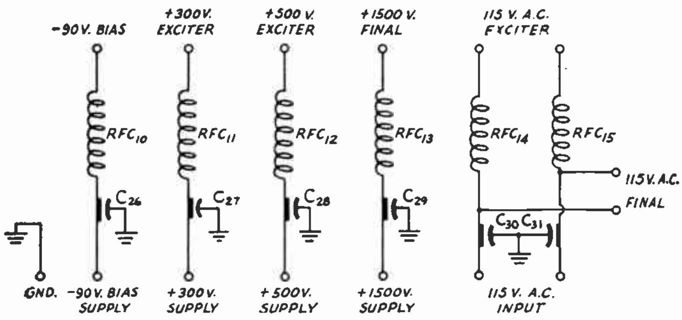


Fig. 6-67 — Wiring diagram of the harmonic-filter unit for the 500-watt all-band transmitter.

C_{26} — 0.005 μfd., 600 volts (Sprague Hypass).
C_{27}, C_{28} — 0.01 μfd., 600 volts (Sprague Hypass).
C_{29} — 0.002 μfd., 5000 volts wkg. (Sprague Hypass).
C_{30}, C_{31} — 0.1 μfd., 250 volts (Sprague Hypass).
RFC_{10-15} — 7-μh. v.h.f. choke (Ohmite Z-50).

readily accessible. Connections to it are made by way of feed-through points in the chassis.

All power wiring is done with shielded conductor and is brought out at the rear to banana-type plugs set in bakelite insulating grommets. R.f. connections should be as short and direct as possible and all by-pass condensers connected to the terminals to be by-passed, and grounded as close as possible to the cathode (or cathode by-pass) grounding point.

Figs. 6-64 and 6-66 show the construction of the final-amplifier section. The chassis is 10 × 17 × 3 inches and is fitted with a standard rack panel of metal 15¾ inches high. The amplifier and antenna tuner are separated by a sheet-aluminum partition. The two tank condensers are mounted on small ceramic cones and placed at an equal distance from the respective ends of the chassis. The coil jack bar in the amplifier is fastened to the tank-condenser frame by means of aluminum angle pieces. The two amplifier tubes are mounted in a manner quite similar to that described for the 807s in the exciter, through clearance holes in the chassis. Clearance holes are also cut for the neutralizing condensers as well as for the tubular condensers, C_{44} and C_{45}. The neutralizing condensers are removed from their original insulator mountings and are fastened instead to large feed-through insulators (Millen type 32103) set in the strip supporting the tube sockets. The connections are then made to the feed-through terminals below. The tubular condensers are placed so that the mounting flanges are close to the point where the filament by-pass condensers are grounded. These condensers are made as shown in Fig. 6-65B.

Underneath, the grid tank condenser is mounted on brackets at the center of the chassis near the strip holding the tube sockets. The brackets space the condenser from the chassis to clear the grid-coil socket which is submounted centrally to the right of the tubes in Fig. 6-64. The two filament transformers are fastened to the right-hand end of the chassis.

A metal strip spanning the antenna tank condenser from front to rear provides a mounting for the antenna tank coil with its axis at right angles to that of the amplifier coil. A Millen right-angle gear box fastened to the partition on metal pillars drives the link-adjustment shaft from the control at the upper center of the panel.

The meter panel is 5¼ inches high. The backs of the meters are shielded by enclosing them in standard metal boxes 3 by 4 by 5 inches.

The frame for the enclosure is made from strips of 1 by 2 stock. Its over-all height (31½ inches) and width (10 inches) match the panel dimensions. The over-all depth is 12 inches. The copper screening is placed on the inside and is brought out around the outer edges of the frame so that it will make an overlapping contact with the metal panels in front and the screening of the removable back. The back extends down only as far as the top edge of the 3 × 4 × 17-inch chassis holding the power-supply terminals. The screening of the hinged top also should make good contact with the screening of the sides. The removable back, the hinged top and the sliding exciter unit are provided with interlock micro-switches that break the power-supply primary circuits when either is opened.

The circuit diagram of a suitable power supply for this transmitter is shown in Fig. 6-69.

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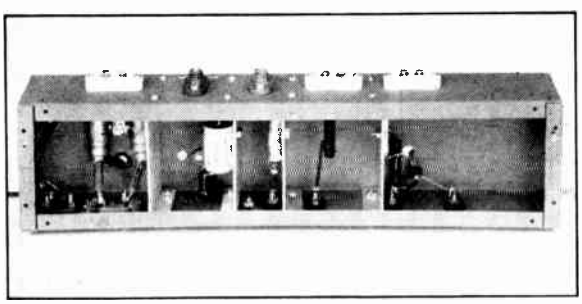


Fig. 6-68 — Bottom view of the line-filter unit. The chassis is divided off into shielded compartments by aluminum partitions. Left to right, the filters are for the a.c. line, 1500-volt d.c. line, 500-volt d.c. line, 300-volt d.c. line, and bias.

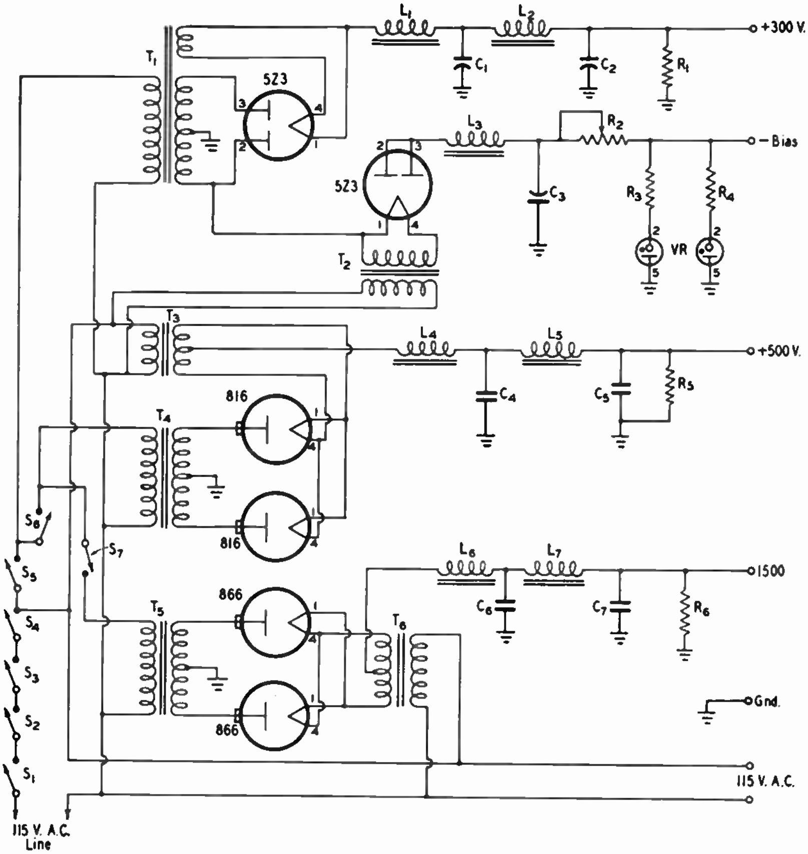


Fig. 6-69 — Circuit diagram of a power supply for the 500-watt all-band transmitter.

C_1, C_2, C_3 — 8-μfd. 450-volt-wkg. electrolytic.
C_4, C_5 — 4-μfd. 600-volt oil-filled.
C_6, C_7 — 4-μfd. 2000-volt oil-filled.
R_1 — 25,000 ohms, 25 watts.
R_2 — 25,000 ohms, 50 watts.
R_3, R_4 — 100 ohms, 1 watt.
R_5 — 25,000 ohms, 100 watts.
R_6 — 25,000 ohms, 150 watts.
L_1, L_2 — 20-hy. 120-ma. filter choke.
L_3 — 20-hy. 75-ma. filter choke.
L_4 — 5/25-hy. 200-ma. swinging choke.
L_5 — 20-hy. 200-ma. smoothing choke.
L_6 — 5/25-hy. 400-ma. swinging choke.
L_7 — 20-hy. 400-ma. smoothing choke.
S_1, S_5, S_6, S_7 — 10-amp. toggle switch.
S_2, S_3, S_4 — Microswitch interlocks.
T_1 — Power transformer: 300 volts d.c., 120 ma.; 5 volts, 3 amp.
T_2 — Filament transformer: 5 volts, 3 amp.
T_3 — Filament transformer: 2.5 volts, 4 amp.
T_4 — Plate transformer: 500 volts d.c., 200 ma.
T_5 — Plate transformer: 1500 volts d.c., 400 ma.
T_6 — Filament transformer: 2.5 volts, 10 amp., 10,000-volt insulation.
VR — VR-90 regulator tube.

Adjustment

The accompanying tables give the coil dimensions and show the coil line-up for any desired output frequency, depending upon VFO or crystal frequency. Care should be taken to check the frequency of each stage with an absorption-type wavemeter until the proper dial settings for each band have been determined and logged. The objective should be to obtain rated grid current to the final amplifier with a minimum of drive to the 807 stage. The coupling between the driver and the final should be adjusted to the optimum point, while the link at the input of the 807s should in each case be set to produce rated final-amplifier grid current.

The grid current to the 6L6 should run 1 ma. or less on all bands. The combined screen and plate current should vary from 10 ma. or less, when the input circuit is untuned, to 45 ma. when the 6L6 is doubling. To obtain rated grid current to a pair of 812As in the final amplifier, as an example, the grid current of a single 807 as a straight amplifier should be about 3 ma. When the two tubes are in use as a doubler, a total grid current of 2 ma. or less should be sufficient. The respective plate currents under these conditions are 100 ma. and 140 ma. The 807 screen current will run between 5 and 7 ma. for single-tube operation and a total of about the same for the two tubes when they are operating as doublers.

When tubes of other types are used in the output stage, R_{12} and R_{13} (Fig. 6-61) must be changed to suit (see "R.F. Power-Amplifier-Tube Operating Factors," this chapter).

A Push-Pull 813 Transmitter

Shown in Figs. 6-70 through 6-77 is an exciter-amplifier combination comprising a complete transmitter capable of 800 watts input in AM 'phone operation and 900 watts in c.w. or NFM service. A pair of 813 beam tetrodes in push-pull is used in the final amplifier. The exciter unit uses an 807 in its output stage and is itself capable of being used as a 75-watt c.w. or 60-watt AM 'phone transmitter. Both units cover all amateur bands from 3.5 to 28 Mc. and are designed for mounting in an enclosed relay rack.

Circuit Details

The diagram of the exciter unit is shown in Fig. 6-71. Provision is made for frequency control from 3.5- and 7-Mc. crystals or from an external VFO unit. Bandswitching is used in all stages except the 807 plate circuit to reduce the number of plug-in coils that must be handled when changing bands, and to permit good isolation between the input and output circuits of the 807. The crystal oscillator uses a 6AG7 in a circuit in which the screen serves as the anode of a Pierce triode oscillator, with the output circuit tunable to either 3.5 or 7 Mc. The tuning condenser, C_3, covers both bands with a single coil in this stage. When VFO control is used, the screen is grounded through C_1 by the crystal-VFO switch and the stage operates as a conventional frequency doubler.

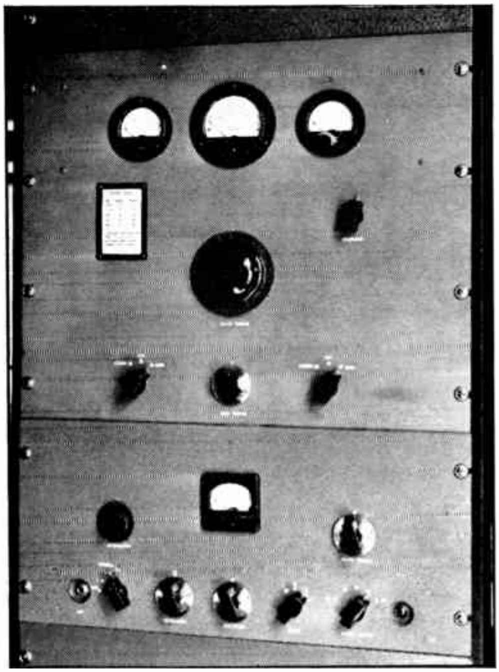


Fig. 6-70 — Front view of the push-pull 813 amplifier and its exciter mounted in a standard rack. From left to right, the controls along the lower edge of the exciter panel are for the crystal-VFO switch, the oscillator tuning condenser, multiplier tuning condenser, meter switch and bandswitch. The controls are flanked by the key jack and a panel lamp. The knob to the left of the exciter milliammeter is the excitation control; the one to the right is the 807 output tuning control.

On the amplifier panel, the meter switches flank the grid tuning dial at the bottom, with the plate tuning dial and the control for the swinging link below the meters.

For either 3.5- or 7-Mc. output, the 6AG7 drives the 807 directly, but for output at higher frequencies, a 6V6 multiplier stage is brought into use by the bandswitch S_2. This stage has two plate coils, L_2 and L_3, the desired one being selected by the bandswitch. One coil is used for the 14-Mc. band, while the other covers both 21 and 28 Mc. The stage operates as a doubler for 14-Mc. output, a tripler for 21 Mc., and as a quadrupler for 28 Mc. The cathode biasing resistor, R_6, protects the tube against excessive input in the absence of excitation.

The 807 stage is operated as a straight amplifier on all bands to reduce harmonics in its output circuit. A 6Y6G is connected as a protective tube to hold the 807 input well below the maximum dissipation rating when excitation is removed. C_{13} is a tubular air condenser connected directly from plate to ground to assist in the reduction of v.h.f. harmonics. In conjunction with RFC_5, it also serves to eliminate high-frequency parasitic oscillation. Plug-in coils are conveniently used in the output circuit of this stage.

The single milliammeter may be switched to read currents essential to the proper tuning of the exciter. All power leads are filtered for v.h.f. harmonics.

The circuit of the push-pull 813 final-amplifier section is shown in Fig. 6-74. The amplifier is link coupled to the exciter. A multiband tuner (National MBS-20) eliminates the need for access to the grid circuit and thus permits complete shielding of the grid circuit for better stability. With this tuner, the grid tank circuit may be resonated anywhere within the frequency range of the transmitter without changing coils.

Small improvised condensers are used to neutralize the amplifier, and chokes inserted in the grid leads eliminate v.h.f. parasitic oscillation.

Three meters are used in the amplifier. One measures the total cathode current of the amplifier, while the others are switched to read individual grid or screen currents of the two tubes, thus permitting a ready comparison of currents for balance in the stage. All supply leads and the leads running to the meters are shielded and filtered to reduce TVI. Plug-in coils are used in the output circuit.

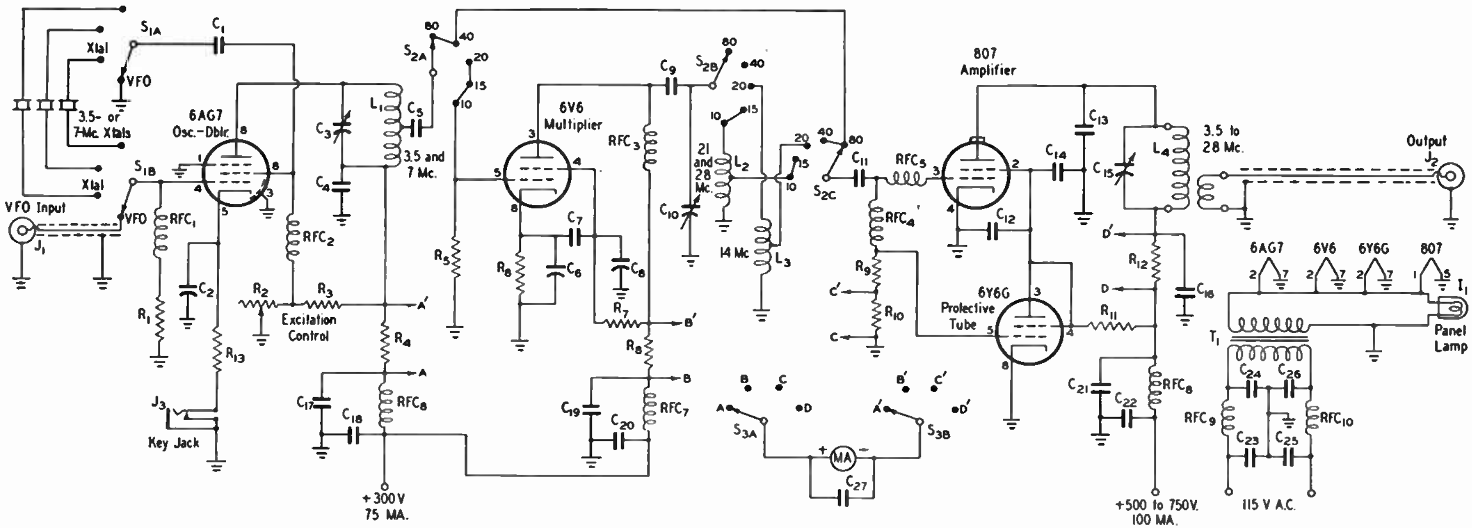


Fig. 6-71 — Circuit diagram of the exciter of the push-pull 813 transmitter.

C_1, C_9, C_{27} — 0.001-μfd. mica.
C_2, C_6, C_8 — 0.01-μfd. paper.
C_3 — 300-μμfd. variable (National STH-300).
C_4, C_{14} — 0.0022-μfd. mica.
C_5, C_{11} — 100-μμfd. mica.
C_7, C_{12} — 22-μμfd. ceramic.
C_{10} — 50-μμfd. variable (National ST-50).
C_{13} — Tubular air condenser, approx. 10 μμfd. (see text).
C_{15} — 150-μμfd. variable, 0.047-inch spacing (National TMK-150).
C_{16} — 0.001-μfd. 1000-volt-wkg. mica.
C_{17}, C_{18}, C_{19}, C_{20}, C_{23}, C_{24}, C_{25}, C_{26} — 470-μμfd. mica.
C_{21}, C_{22} — 500-μμfd. 1000-volt-wkg. mica.
R_1 — 15,000 ohms, ½ watt.
R_2 — 25,000-ohm 7-watt wire-wound potentiometer.
R_3 — 17,500 ohms, 10 watts.
R_4, R_8, R_{10} — 100 ohms, ½ watt.
R_5 — 56,000 ohms, 1 watt.
R_6 — 600 ohms, 2 watts.
R_7 — 25,000 ohms, 10 watts.
R_9 — 22,000 ohms, 1 watt.
R_{11} — 50,000 ohms, 10 watts.
R_{12} — 3-times meter shunt (see Chapter 16).
R_{13} — 330 ohms, 1 watt.
L_1 — 22 turns No. 20, 1-inch diam., 1⅜ inches long, tapped 7 turns from plate end (B & W 3015 Miniductor).
L_2 — 8 turns No. 18, ¾-inch diam., 1 inch long, tapped 2 turns from plate end (B & W 3010 Miniductor).
L_3 — 11 turns No. 20, ¾ inch long, ¾-inch diam., tapped 3 turns from plate end (B & W 3011 Miniductor).
L_4 — Millen 43000 series coils, modified:
— 3.5 Mc. — 22 turns No. 20, 1½ inches diam., 1⅛ inches long, 5-turn link (Millen 43082, 18 turns removed).
— 7 Mc. — 14 turns No. 18, 1½ inches diam., 1¼ inches long, 7-turn link (Millen 43012, 8 turns removed).
— 14 & 21 Mc. — 5 turns No. 18, 1½ inches diam., ¾ inch long, 2-turn link (Millen 43022, 4 turns removed).
— 28 Mc. — 3 turns No. 18, 1½ inches diam., 1 inch long, 2-turn link (Millen 43012, 1 turn removed).
I_1 — Panel lamp
J_1, J_2 — Coaxial connector.
J_3 — Closed-circuit jack.
MA — 50-ma. d.c. milliammeter.
RFC_1, RFC_2, RFC_3, RFC_4 — 2.5-mh. r.f. choke.
RFC_5 — 1-μh. r.f. choke (National R-33).
RFC_6, RFC_7, RFC_8, RFC_9, RFC_{10} — 7-μh. r.f. choke (Ohmite Z-50).
S_1 — Two-section ceramic rotary switch, points per deck optional.
S_2 — Three-section 5-position ceramic rotary switch.
S_3 — Two-section 4-position rotary switch.
T_1 — Filament transformer: 6.3 volts, 4 amp. (Thordarson T2F11).

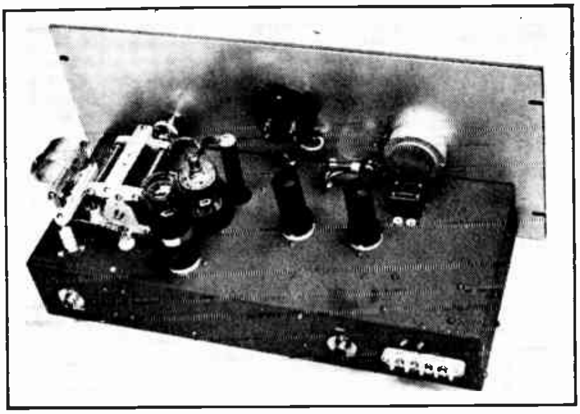


Fig. 6-72 — Rear view of the exciter chassis for the push-pull 813 transmitter. The tubular condenser is alongside the 807 in back of the 6Y6G. The 6V6 multiplier is in the center of the chassis, with the 6AG7 oscillator, the crystal sockets and the excitation control at the right. Coaxial connectors for output and VFO input, plus a terminal strip, are mounted on the rear edge of the chassis.

Construction

The exciter unit is built on a 17 × 7 × 3-inch chassis. The key jack, S_1, C_3, C_{10}, meter switch, bandswitch shaft and the panel lamp are first arranged so that the controls will be equally spaced along the lower edge of the panel. C_3 is insulated by mounting it on a polystyrene subpanel and using an insulating shaft coupling. The similar subpanel holding C_{10} in place is of metal. The crystal sockets are lined up behind the crystal switch and the sockets for the 6AG7 and 6V6 are placed at the center of the chassis alongside their tank condensers.

A Millen type 80070 shield-and-bracket assembly is used for the 807. The tubular condenser, C_{13}, which is similar to the one shown in Fig. 6-65A, is mounted at one corner of the bracket, a hole being cut in the bracket to clear the bottom ceramic button. Clearance holes for the tube shield and the tubular condenser are cut in the chassis so that the bracket can be centered between the front and back edges of the chassis with the bottom of the bracket two inches below the surface of the chassis. It is held in this position by an aluminum-sheet bracket 5 inches long and 2¾ inches deep fastened to the chassis. The socket for the 6Y6G is placed immediately to the rear of the 807.

The bandswitch is fastened, about centrally, on brackets along the rear edge of the chassis. The gear drive (National ACD-2), also on brackets, is lined up with the switch shaft and the panel control.

On top, the output tank condenser and coil are placed close to the 807. The condenser is mounted on ceramic button insulators with its stator terminals on top so that the plate lead can be made short. An insulating coupling is used in the control shaft. The coil socket is elevated on 1-inch ceramic pillars (National GS-1). The excitation control, R_2, is mounted on the panel so as to balance the output tank-condenser tuning control.

The power wiring should be done before the assembly has progressed too far. Shielded wire, laid close to the chassis, should be used. The shielding should be grounded at each end of each lead and at intermediate points where mounting screws, or other grounded metal, make it convenient. Wherever wires cross or run parallel, they should be spot-soldered together. At points where there is danger of a short circuit by the braid, the wire may be covered with a sleeve of spaghetti.

By-pass condensers should be connected with leads as short as possible. Two screen by-pass condensers are shown in the 807 circuit. One of these, C_{12}, is a small ceramic unit soldered directly between the screen and cathode terminals of the socket to serve as a low-inductance path for v.h.f. The other, C_{14}, is grounded at one end on one of the socket-mounting screws. One end of the parasitic suppressor, RFC_5, should be soldered directly to the grid terminal of the socket. The cathode should be grounded with a short lead to the mounting flange of the tubular condenser.

Coaxial connectors are provided at the rear for VFO input and r.f. output to the final amplifier. A terminal strip is set in the rear of the chassis, at the left-hand end in Fig. 6-73, for power-supply connections. The v.h.f. filter components are assembled on a terminal board placed close to this terminal strip. The filament transformer is immediately behind, fastened to the end of the chassis.

After the power wiring has been done, the exciter coils may be put in place. L_1 is mounted on ½-inch cone insulators to the rear of C_3. L_2 and L_3 are placed at right angles (L_2 horizontal and L_3 vertical) behind C_{10} and are soldered between the rotor terminal of the condenser and the S_{2B} section of the bandswitch.

The final amplifier is assembled on a 17 × 13 × 3-inch chassis with a 17½-inch metal panel. The tank condenser is mounted at the exact center of the chassis on 1-inch cone insulators. A high-voltage insulating coupling is placed between the condenser shaft and the control on the panel. The fixed condenser, C_2, is placed under the condenser frame and is connected between the frame and a grounding screw in the chassis. This screw also is used for grounding the grid tuner below.

Clearance holes are cut in the chassis and the sockets are submounted on 1½-inch spacers so that the plate caps of the two tubes will come close to the outside terminals of the condenser stators. A large feed-through insulator is placed 1½ inches from the inside edge of each of the clearance holes. A ½-inch strip of aluminum, about 2½ inches long, is bent into "L" shape and mounted on top of each feed-through. This serves as one side of the neutralizing condenser, the plate of the tube itself forming the other side of the condenser.

To the rear of the tank condenser, the coil jack bar is mounted on large stand-off insulators (National GS-4) to bring the coil terminals close to those of the tank condenser. The link is adjusted from the panel by means of a right-angle gear drive (National ACD-2) mounted from a bracket fastened at a rear corner of the chassis.

The three meters are enclosed in a standard 3 × 4 × 17-inch chassis acting as a shielding box. The box is fastened to the panel with self-tapping screws. Standard 10-inch panel brackets are fastened to the ends of the meter box as well as to the panel and chassis. Power terminals and connectors for r.f. input and output are lined up along the rear edge of the chassis.

Underneath, the grid tuner is mounted at the center of the chassis on pillars to space the coils equally between the chassis and its bottom plate. The individual filament transformers are placed close to their associated sockets. The lower terminals of the two feed-through insulators are connected to opposite (not adjacent) grid terminals. One end of the parasitic-suppressor chokes is soldered directly to the grid terminal of the socket. A 1-inch ceramic pillar at the forward inside corner of each tube socket serves as an insulated tie point for the parasitic choke, the grid choke, the fixed grid condenser and the neutralizing lead on each side of the circuit.

A terminal board at the rear holds the v.h.f. filter components for the a.c. and bias lines. Filters in the other power leads are placed close to their respective terminals. All power wiring is done with shielded wire. The high-voltage lead is a piece of ignition cable covered with a sheathing of copper braid. Shielded leads also connect the meter switches underneath the chassis to the meters on the panel above. C_{24} and C_{25} are connected directly across the terminals of the meters, but RFC_{14} and RFC_{15} are placed under the chassis at the switch terminals.

Adjustment

Fig. 6-77 shows the circuit diagram of a power-supply system for this transmitter. The section at the bottom supplies low voltage for the exciter and bias for the final amplifier, while the next section above supplies voltage for the 807 driver and screen voltage for the output stage. Starting at maximum resistance, R_3 is adjusted until at least one of the VR tubes just ignites. R_4 need not be used, or may be shorted out, for c.w. operation. For plate modulation at maximum ratings, R_4 should be set at 830 ohms. When S_6 is open, reduced screen voltage is applied to the 813s. With S_6 closed, R_7 should be set at approximately 1250 ohms for a supply voltage of 500, or at about 9400 ohms if the supply delivers 750 volts, with proportionate values for voltage between these extremes. After the final amplifier has been adjusted for operation at full load, R_7 should be adjusted finally to bring the screen voltage to 400 for c.w. or 350 for 'phone under operating conditions. S_6 should always be open during preliminary adjustments of the final amplifier or regular adjustment of the exciter, since full screen voltage in the absence of plate voltage and full load can cause dangerous heating of the screen.

The power switches are arranged in series so that the lower voltages must be turned on before the higher voltages can be applied. Under normal operating conditions, all switches will be closed except S_2 which then serves as the power control for the entire transmitter.

The exciter should be tuned up initially with an absorption wavemeter to make certain

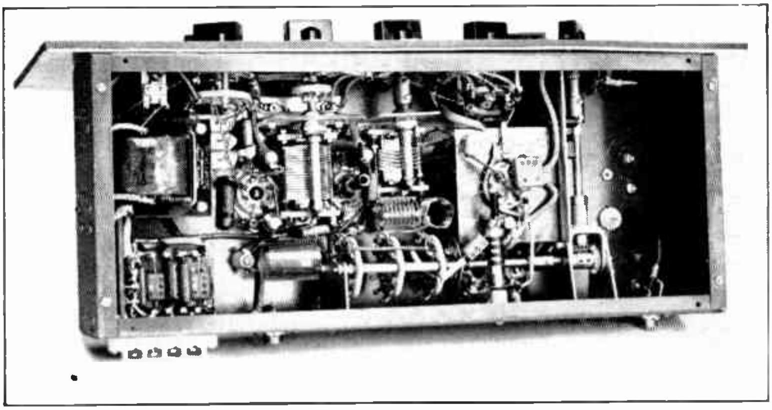


Fig. 6-73 — Bottom view of the exciter for the push-pull 813 transmitter. The harmonic filters are mounted on terminal boards placed in the lower left-hand corner, adjacent to the input terminals and just below the filament transformer. The mounting brackets for the bandswitch and the right-angle drive are supported by the rear of the chassis, while the bracket that supports the 807 socket assembly extends back from the front.

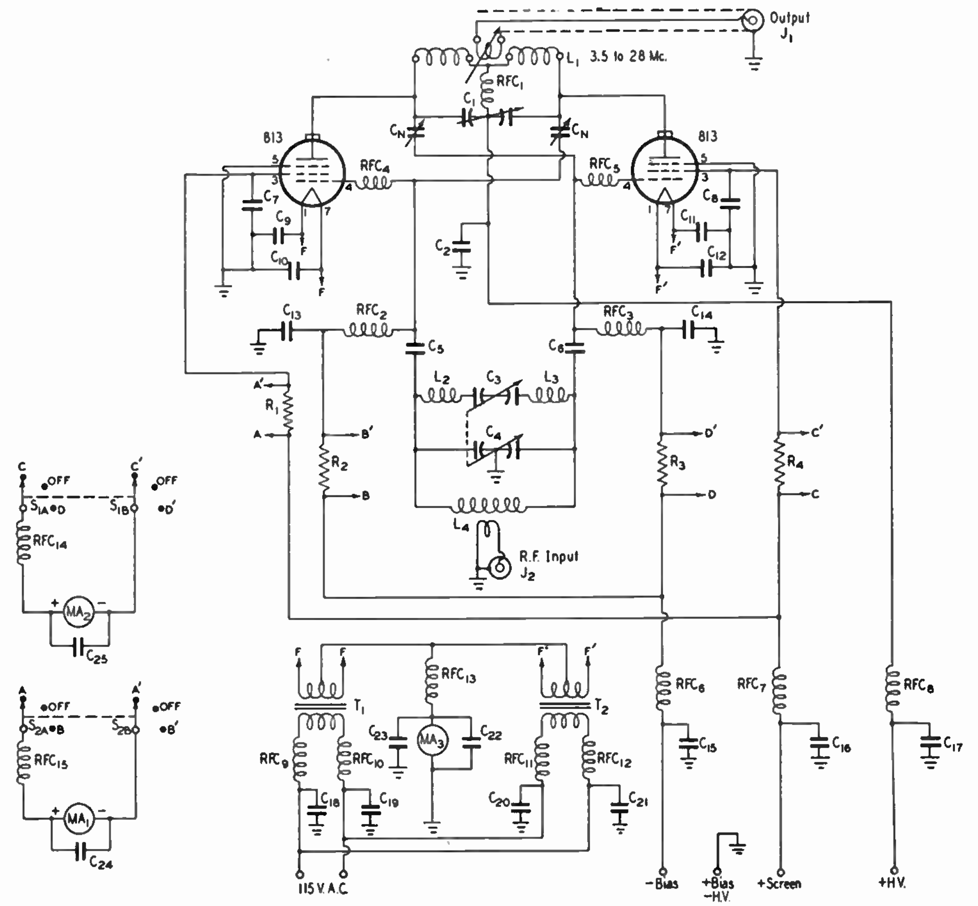


Fig. 6-74 — Schematic diagram of the push-pull 813 amplifier.

C_1 — 100-μμfd.-per-section variable, 0.077-inch spacing (Millen 04103).
C_2 — 0.001-μfd. 5000-volt-wkg. mica.
C_3, C_4 — 125-μμfd.-per-section variable, 0.026-inch spacing (National SSH-125 — part of MB-20 tuner).
C_5, C_6 — 0.001-μfd. mica.
C_7, C_8 — 0.001-μfd. 1000-volt-wkg. mica.
C_9, C_{10}, C_{11}, C_{12} — 0.0047-μfd. mica.
C_{13}, C_{14} — 100-μμfd. mica.
C_{15}, C_{18}, C_{19}, C_{20}, C_{21}, C_{22}, C_{23}, C_{24}, C_{25} — 470-μμfd. mica.
C_{16} — 500-μμfd. 1000-volt-wkg. mica.
C_{17} — 500-μμfd. 5000-volt-wkg. mica.
C_N — See text.
R_1, R_2, R_3, R_4 — 100 ohms, ½ watt.
L_1 — B & W HDVL series coils:
(All are split-winding coils, ¾ inch between sections for all except 21- and 28-Mc. coils where the spacing is 1¼ inches. Dimensions given are for each section of coil.)
— 3.5 Mc. — 16 turns No. 10, 3½ inches diam., 3 inches long.
— 7 Mc. — 10 turns No. 8, 3½ inches diam., 2⅞ inches long.
— 14 Mc. — 6 turns No. 8, 3½ inches diam., 3 inches long.
— 21 Mc. — 4 turns 3/16-inch copper tubing, 3 inches diam., 2⅞ inches long.
— 28 Mc. — 2 turns 3/16-inch copper tubing, 3 inches diam., 2⅞ inches long. (One turn removed from each section of HDVL-10).
L_2, L_3 — 7 turns No. 22, 1-inch diam., 5/16 inch long, ⅜ inch between windings (part of MB-20 tuner).
L_4 — 30 turns No. 22, 1-inch diam., 1¼ inches long (part of MB-20 tuner).
J_1, J_2 — Coaxial connector.
MA_1, MA_2 — 50-ma. d.c. milliammeter.
MA_3 — 750-ma. d.c. milliammeter.
RFC_1 — 800-ma. r.f. choke (National R-175).
RFC_2, RFC_3 — 2.5-mh. r.f. choke.
RFC_4, RFC_5 — 1-μh. 300-ma. r.f. choke (National R33).
RFC_6, RFC_7, RFC_8, RFC_9, RFC_{10}, RFC_{11}, RFC_{12}, RFC_{13}, RFC_{14}, RFC_{15} — 7-μh. r.f. choke (Ohmite Z-50).
S_1, S_2 — 2-section 3-position ceramic rotary.
T_1, T_2 — Filament transformer: 10 volts, 5 amp. (Thordarson T21F18).

that each circuit is tuning to the proper frequency. At the plate voltage specified, the meter reading with S_3 in the first position should run between 25 and 35 ma., depending upon the setting of the excitation control, R_2. The combined screen and plate current of the 6V6 before excitation is applied should be approximately 30 ma., increasing to 35 ma. when excitation is applied and the stage is doubling or tripling, and to about 45 ma. when quadrupling. The grid current to the 807 should be adjusted, by means of the excitation

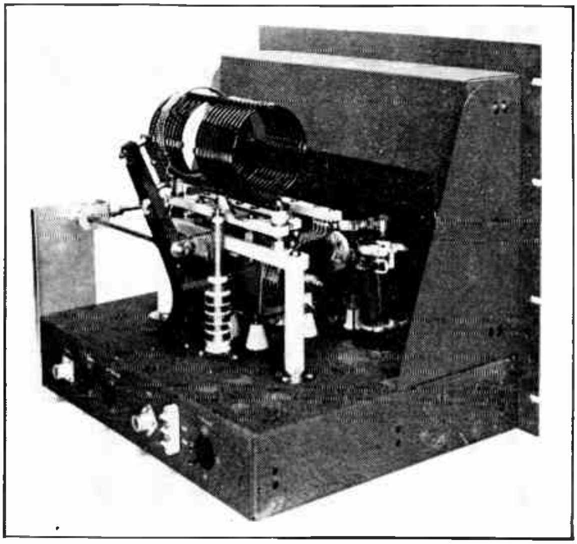


Fig. 6-75 — Rear view of the push-pull 813 amplifier. The feed-through insulator holding one of the neutralizing condensers is just to the left of the visible 813. The chassis that encloses the meters is held in position with self-tapping screws passing through the up-ended panel brackets. The gear drive at the left is for link adjustment from the panel. Input, output and all power connections are arranged along the rear edge of the chassis.

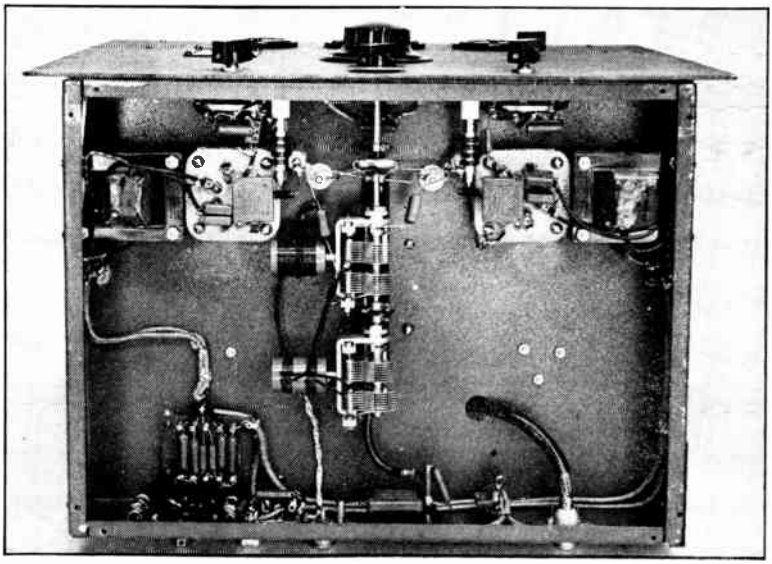
control, to the minimum that will give rated grid current to the final amplifier with optimum coupling to the 807 when the final is loaded fully. It should be possible to do this without exceeding a grid current of 3.5 ma. to the 807 and with the plate current between 60 and 90 ma.

In adjusting the multiband tuner in the amplifier grid circuit, the resonances should be checked carefully with an absorption wavemeter to make sure that the circuit is tuned to the desired frequency. The setting of each band should then be logged.

Since the adjustment is more critical at 28 Mc. than on any of the other bands, the final stage should be neutralized with the transmitter tuned to this band. With an indicating absorption wavemeter or other r.f. indicator coupled to the output tank coil and with excitation only applied, the grid and plate tank circuits should be tuned to resonance. Resonance in the output tank circuit will be indicated by a maximum response on the indicator. The neutralizing condensers should then be adjusted similarly, bit by bit, either by bending the metal strips closer to, or farther away from, the tubes, or by clipping the length of the strips until a minimum response on the neutralizing indicator is obtained when the plate tank circuit is tuned to resonance. In this particular amplifier, minimum r.f. feed-through was obtained with the strips clipped to about a half inch.

To check the balance of the amplifier, temporarily disconnect the two center-tap leads of the filament transformers from the cathode meter and insert individual meters between the center taps and ground. Apply power to the exciter with the transmitter tuned to the 28-Mc. band. Resonate the grid circuit and set the meter switch to read individual grid currents. The readings may not be equal before plate and screen voltages are applied to the final amplifier, but the readings should rise and fall together as the grid circuit is tuned through

Fig. 6-76 — Bottom view of the push-pull 813 amplifier. The multiband tuner used in the grid circuit is centrally located, flanked by the two tube sockets. All by-pass condensers are mounted on the sockets. The neutralizing leads are crossed beneath the insulated shaft coupling, and terminate at stand-off insulators placed close to the grid terminals of the tube sockets. The harmonic filters are placed along the edge of the chassis close to the points at which the various leads leave the chassis. The coaxial cable to the right is the output link line.



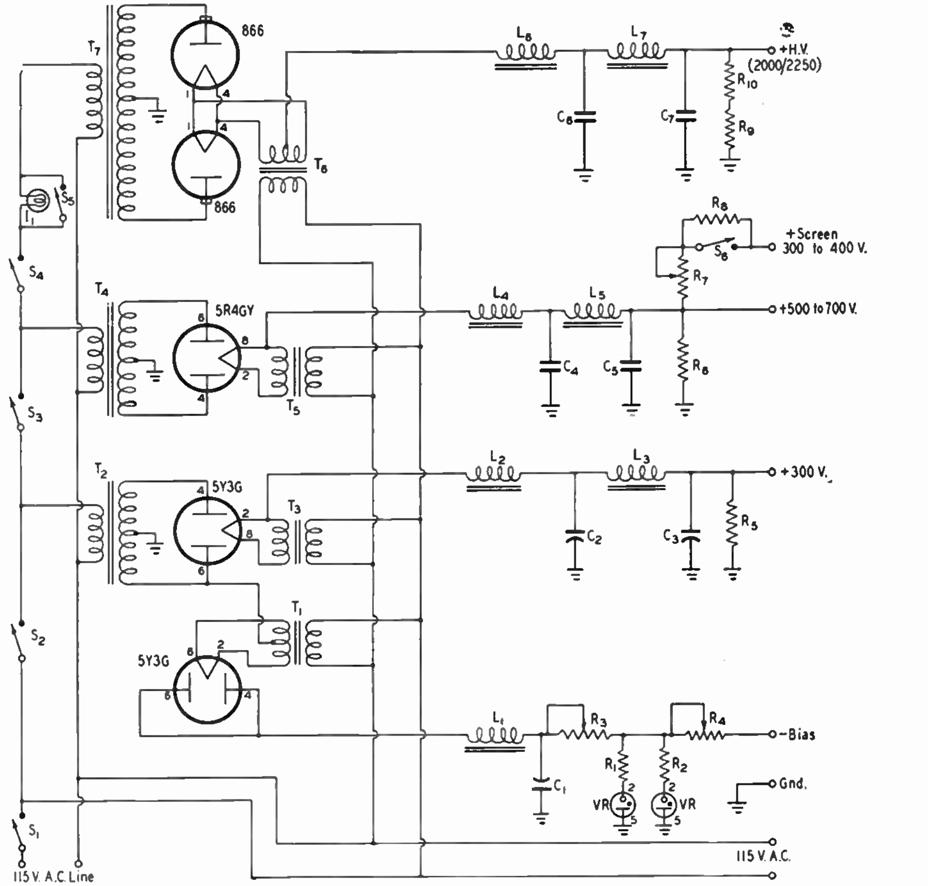


Fig. 6-77 — Circuit diagram of a power-supply system for the push-pull 813 transmitter.

C_1, C_2, C_3 — 8-μfd. 150-volt electrolytic.
C_4, C_5 — 4-μfd. 1000-volt oil-filled.
C_6 — 2-μfd. 2500-volt oil-filled.
C_7 — 4-μfd. 2500-volt oil-filled.
R_1, R_2 — 100 ohms, 1 watt.
R_3 — 25,000 ohms. 25 watts, adjustable.
R_4 — 1000 ohms, 10 watts, adjustable.
R_5 — 15,000 ohms, 10 watts.
R_6 — 25,000 ohms, 50 watts.
R_7 — 10,000 ohms, 50 watts, adjustable.
R_8 — 10,000 ohms, 50 watts.
R_9, R_{10} — 25,000 ohms, 75 watts.
L_1 — 30-hy. 50-ma. filter choke.
L_2, L_3 — 20-hy. 100-ma. filter choke.
L_4 — 5/25-hy. 150-ma. swinging choke.
L_5 — 20-hy. 150-ma. smoothing choke.
L_6 — 5/25-hy. 500-ma. swinging choke.
L_7 — 20-hy. 500-ma. smoothing choke.
I_1 — 115-volt lamp of suitable size to reduce voltage for tune-up.
S_1 — 20-amp. s.p.s.t. switch.
S_2, S_3, S_4 — 15-amp. s.p.s.t. switch.
S_5 — 10-amp. s.p.s.t. switch.
S_6 — Ceramic s.p.s.t. rotary switch.
T_1, T_3, T_5 — Filament transformer: 5 volts, 3 amp.
T_2 — Power transformer: 150 volts r.m.s. each side of center, 100 ma.
T_4 — Plate transformer: 500 to 750 volts d.c., 150 ma.
T_6 — Filament transformer: 2.5 volts, 10 amp., 10,000-volt insulation.
T_7 — Plate transformer: 2000/2250 volts d.c., 500 ma.
VR — VR-150-30.

resonance. If such is not the case, a slight readjustment of the position of the grid link should improve this condition. In some cases it may be necessary to connect a small padding condenser across one of the two sections of C_3 and adjust it until the grid currents rise and fall in unison and are reasonably well balanced.

With a dummy load connected to the output, apply reduced screen and plate voltages, resonate the tank circuits and observe grid, screen and cathode currents of the two tubes. An agreement within 10 per cent may be considered satisfactory. If the difference is greater, check the wiring in the plate circuit to be sure that it is symmetrical. A slight difference in lead length, between the tank circuit and the tubes, can cause considerable unbalance at 28 Mc. Some readjustment of the grid padding condenser, if one is used, may help.

In c.w. service, plate voltages up to 2250 may be used and up to 2000 for AM 'phone. Maximum plate current under these conditions should be 220 and 200 ma. respectively per tube. The total of grid and screen currents of both tubes must be subtracted from the reading of the cathode meter to obtain the actual plate current. Screen current should be less than 40 ma. per tube under full load.

A High-Stability VFO

Figs. 6-78 through 6-84 illustrate the construction of a well-stabilized VFO delivering sufficient output at 1.75, 3.5 or 7 Mc. to drive any small triode or tetrode stage up to and including one or two 807s. The two output frequences are chosen so that if the VFO is used to drive a crystal-oscillator stage in the transmitter, the oscillator stage may be operated as a doubler to avoid the possibility of oscillation in the crystal stage.

Referring to the diagram of Fig. 6-79, a 6AG7 pentode is used in the electron-coupled series-tuned Colpitts oscillator circuit. C_1 is the bandspread tuning condenser which covers the fundamental range of 1750 to 2000 kc. (or 3500 to 4000 kc.). C_2 is a padder to provide a fixed minimum circuit capacitance. C_3 and C_4 are the tube-shunting capacitances. Since the screen, which serves as the plate in the oscillating circuit, is grounded, the cathode is above ground and therefore must be returned to ground for d.c. through a choke.

The output circuit (RFC_2) is nonresonant and is capacity coupled to a 6L6 output stage fitted with plug-in coils so that it may be operated at either 1.75, 3.5 or 7 Mc. The tuning condensers of the oscillator and amplifier are ganged.

A power supply is included in the unit. Screen and plate voltages for both stages are taken from a VR-tube voltage divider. The regulator tubes are used both as a convenient voltage-divider arrangement and to limit the shaping of the keying characteristic entirely to any key-click filter that may be used with the unit.

Construction

In the unit shown in the photographs, the frequency-determining tank is isolated from the rest of the circuit by enclosing it in a standard steel box 5 × 6 × 9 inches. The tuning condenser is mounted on the top plate of a 4 × 4 × 2-inch steel box with metal brackets that space the bottom edges of the condenser end plates 1/8 inch from the plate.

The coil is removed from its original mounting, the link removed, and the coil remounted on a 3/4-inch cone insulator at the forward end and a small feed-through insulator at the rear. The first quarter turn at the front end of the coil is broken loose and a short connection between the adjacent condenser terminal and the coil at this point is made with a piece of heavy wire. This serves as a brace for the coil against vibration. Another short piece of heavy wire goes from this point to a small feed-through insulator set directly below in the top plate. This feed-through insulator and the one at the rear end of the coil serve in making connections to the condensers on the underside of the plate.

The adjustable padder, C_2, is mounted centrally on the underside of the plate with its shaft pointed toward the right. The end of the shaft is slotted for a screwdriver and holes

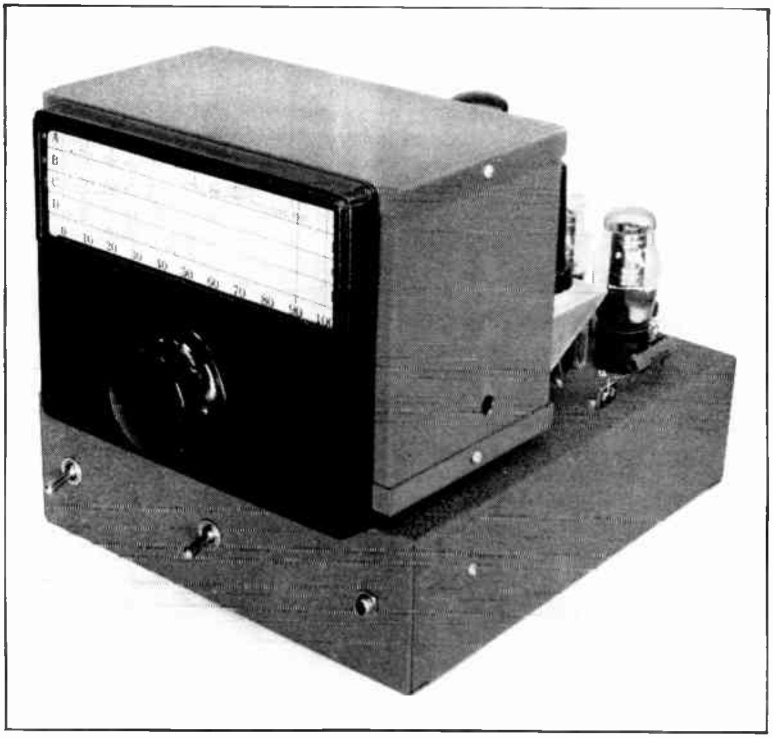


Fig. 6-78 — The completed VFO. The entire r.f. section is floating on an anti-shock mounting. The hole in the side of the box is to permit adjustment of the frequency range

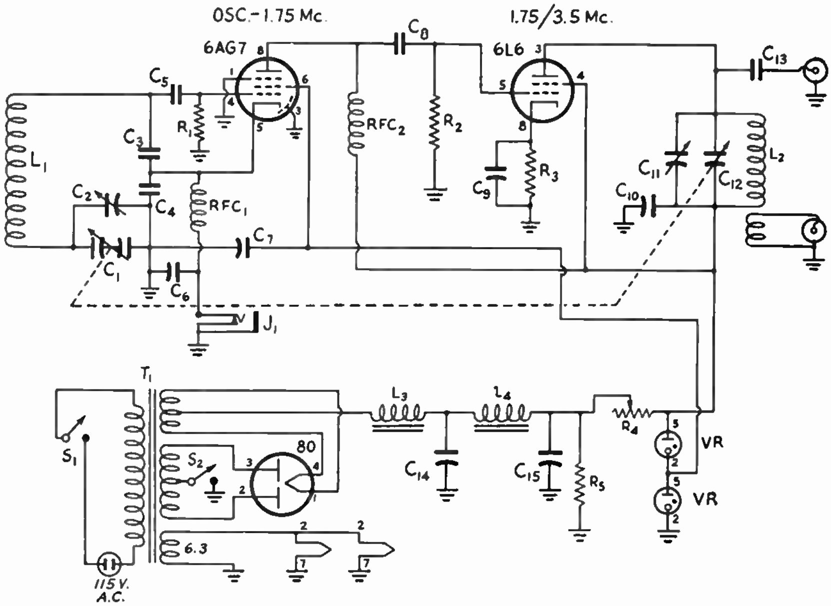


Fig. 6-79 — Circuit diagram of the series-tuned VFO.

C_1 — 50-μμfd.-per-section variable (Millen 23050).
C_2 — 100-μμfd. variable (Millen 19100).
C_3, C_4 — 0.001-μfd. zero-temp. mica.
C_5, C_8 — 100-μμfd. mica.
C_6, C_7, C_9, C_{10} — 0.01-μfd. paper.
C_{11} — 1.75 and 3.5 Mc. — 45-260-μμfd. mica trimmer.
7 Mc. — 100-μμfd. air trimmer (Hammarlund APC-100).
C_{12} — Approx. 75-μμfd. variable (Millen 22100 with 3 stator plates removed).
C_{13} — 220-μμfd. mica.
C_{14}, C_{15} — 16-μfd. 450-volt electrolytic.
R_1 — 47,000 ohms, ½ watt.
R_2 — 0.1 megohm, 1 watt.
R_3 — 470 ohms, 1 watt.
R_4 — 1000 ohms, 10 watts, adjustable.
R_5 — 50,000 ohms, 10 watts.
L_1 — 1.75 Mc. — 140 μh. (National AR-160).
3.5 Mc. — 35 μh. (B & W JEL-80).
L_2 — 1.75 Mc. — 37 turns No. 22 d.c.c., 1½ inches diam., close-wound.
3.5 Mc. — 16 turns No. 22, 1½ inches diam., ⅞ inch long.
7 Mc. — 14 turns No. 20, 1½ inches diam., 1½ inches long, tapped 5½ turns from ground end for C_{11}.
L_3, L_4 — 11-h. 100-ma. filter choke (UTC R-19).
J_1 — Closed-circuit jack.
RFC_1, RFC_2 — 2.5-mh. r.f. choke.
S_1, S_2 — S.p.s.t. toggle switch.
T_1 — Power transformer: 350-350 r.m.s., 90 ma.; 5 volts, 3 amp.; 6.3 volts, 3.5 amp.

are drilled in the sides of both inner and outer boxes so that the padder may be adjusted from the outside after the unit has been assembled. The mica condensers, C_3 and C_4, are fastened alongside the padding condenser by cementing them to the plate with Duco cement to eliminate movement. The top lip of the small box may have to be notched out in a few places before the top plate will fit in place.

Discarding the bottom plate of the small box, the height

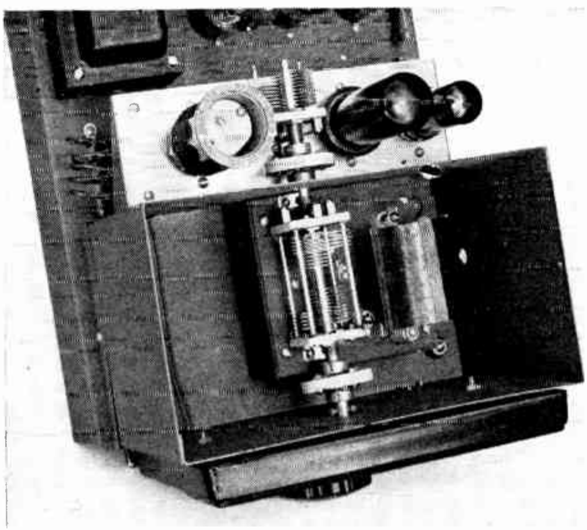


Fig. 6-80 — The oscillator tank circuit is isolated from the remainder of the circuit by enclosing it in a shock-proof metal box.

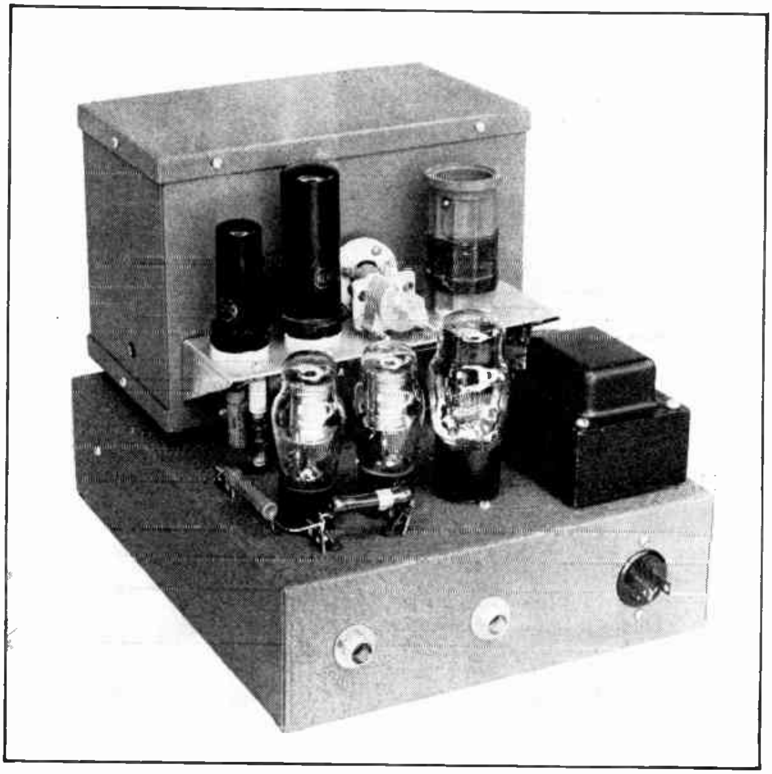


Fig. 6-81 — Rear view of the VFO. The oscillator tube and amplifier components are mounted on a shelf fastened to the floating box. Power-supply components occupy the rear portion of the chassis.

of the tuning-condenser shaft above the lower edge of the box should be measured carefully and 1-inch clearance holes cut centrally in the outer box at this same level. Placing the smaller box inside, with its rear face against the back wall of the outer box and with the tuning-condenser shaft lined up with the shaft holes, the position of the smaller box should be marked on the rear wall. Then the top plate should be removed, the small box replaced, and holes marked in the bottom of the outer box so that the smaller box can be fastened in place with screws up through the bottom. With this done, a grommet hole for the leads to the oscillator tube should be drilled through the rear of both boxes simultaneously near the oscillator-tube socket. Three leads — connections to the grid condenser, C_5, to the cathode, and to the ground point of the cathode by-pass condenser of the oscillator tube — are bunched together and brought out through this hole.

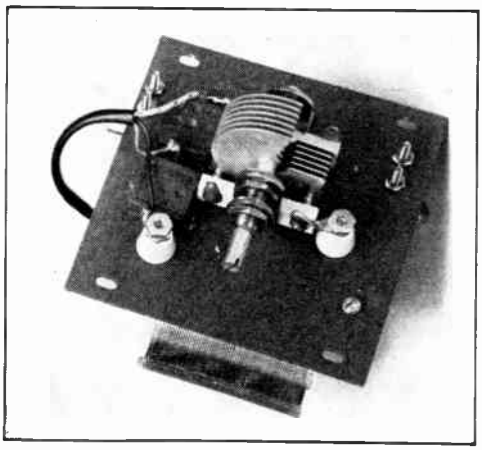


Fig. 6-82 — Bottom view of the oscillator-tank unit, showing the placement of the oscillator tuning padder and the tube-shunting condensers.

With the oscillator-tank unit fastened in place within the large box, and flexible insulated couplings on each end of the tuning-condenser shaft, the dial can be lined up and its mounting holes marked on the front of the outer box. The lower edge of the dial plate will overhang a half inch or so at the bottom.

The remainder of the r.f.-circuit components are assembled on a $2\frac{1}{2} \times 8$-inch aluminum shelf fastened to the rear of the box to isolate the tank components from the heating of the tubes. The amplifier tuning condenser, C_{12}, must be insulated from the shelf. The height of the shelf is adjusted, after the condenser has been mounted, so that its shaft lines up with the tail shaft of C_2. Wiring and associated small parts are placed under the shelf. All power-supply connections and the key connection are made to a 5-point lug strip at the left-hand end of the shelf.

The entire unit is guarded against mechanical vibration by mounting the box on rubber grommets. A grommet is placed in each of the four corners of the bottom of the

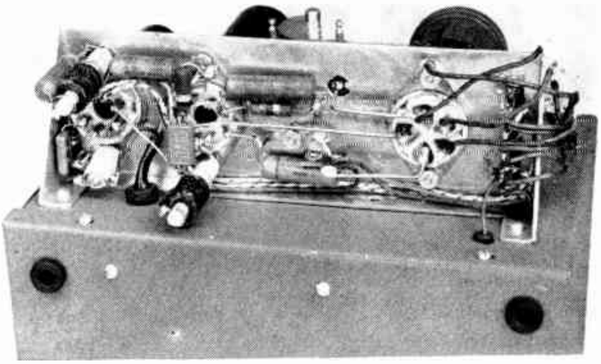


Fig. 6-83 — Wiring underneath the shelf. The loose flexible leads at the right are anchored to a lug strip on the chassis after assembly.

box. These are duplicated in the top of the 10 × 12 × 3-inch chassis which serves as a base. Machine screws with washers at either end are passed through both sets of grommets to fasten the floating box to the chassis. Care should be taken in locating the grommet holes in the chassis to provide $\frac{1}{16}$ inch or so of clearance between the lower overhanging edge of the dial plate and the chassis, so that the dial is free from contact with the chassis.

A duplicate lug strip is fastened to the chassis directly below the terminal strip on the shelf. The two strips are then connected together with highly-flexible wire bent to form half loops between the terminal strips. This is done to minimize any vibration that might be transmitted from the base chassis to the box through the connecting leads. Similar flexible connections are made to anchorages on the chassis for the output leads.

The output coil, L_2, is wound on a standard 1½-inch diameter 5-prong plug-in form (Bud). The padder condenser, C_{11}, in each case is mounted inside the form where it may be adjusted with a screwdriver.

The power transformer, rectifier, the two VR tubes and their voltage-dropping resistor, R_4, as well as the bleeder resistor, R_5, are mounted along the rear edge of the chassis. The filter chokes and condensers are placed underneath, since they develop no appreciable heat. A 115-volt connector and two coaxial output connectors are mounted in the rear edge of the chassis. The output may be either capacity or link coupled to a following stage. The two power switches and the key jack are set in the front edge of the chassis.

Adjustment

The adjustment of the unit is very simple. The VR resistor, R_4, should first be set so that the VR tubes stay ignited with the key closed. VR75s or VR90s may be used, the higher voltage giving somewhat greater output from the unit. Then, the tuning condenser C_1 should be set at maximum capacitance. Listening on a receiver tuned to about 3490 kc., the oscillator padder, C_2, should be adjusted until the oscillator signal is heard at that frequency. The oscillator tuning should then cover the range up to a frequency slightly above 4000 kc.

The amplifier padder is adjusted by tuning the oscillator to the approximate center of the band and adjusting C_{11} for maximum grid current to the following stage. If the coil dimensions have been followed carefully, the output should then be substantially constant over the entire band. The 1750-kc. or 3.5-Mc. output should be used in feeding a crystal stage normally using 3.5-Mc. crystals, while 3.5- or 7-Mc. output should be used in cases where 7-Mc. crystals are normally employed.

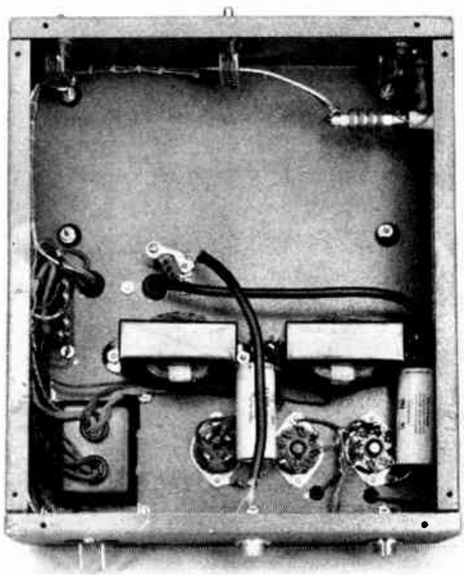


Fig. 6-84 — Bottom view of the completed series-tuned VFO unit. Power leads are cabled. The coaxial cables go to the r.f. output connectors.

A Bandswitching Gang-Tuned VFO-Exciter

Figs. 6-85 through 6-88 show the construction of a bandswitching gang-tuned VFO-exciter delivering an output of approximately 5 watts on all bands from 80 to 10 meters. Power supply and a reactance modulator for NFM are included.

Referring to the circuit diagram of Fig. 6-87, a 6AG7 is used in the series-tuned Colpitts oscillator circuit. Three frequency ranges are covered here. The first range, with L_1 in the circuit, is from 3.5 to 4 Mc. This range is in use for 3.5-, 7- and 14-Mc. output from the exciter. The second range, covered by L_2, starts at 5.25 Mc. to cover the 21-Mc. band in the output. The third range, with L_3, starts at 6740 kc. to cover the 11- and 10-meter bands in the output.

The second stage is a 6L6 buffer-doubler. In the 80- and 40-meter positions of the bandswitch, the output circuit is resonant approximately halfway between the 80- and 40-meter bands. This provides sufficient excitation for the output stage without danger of instability. In the last three positions of the switch, the stage doubles frequency successively to 7, 10.5 and 14 Mc. The last stage operates as a straight amplifier with an untuned input circuit on 80 meters, and as a doubler to 7, 14, 21 and 27–28 Mc. The tuning condensers of all three stages are ganged.

The reactance modulator consists of a 6AK5 amplifier driving a 6BA6 reactance tube. Jacks are provided for keying either the oscillator or the output stage, as desired. The screen voltage of all three r.f. tubes, as well as the plate supply for all but the output tube, is regulated by VR tubes. S_1 is a control switch. In the first position, all power is turned off. In the second position the power supply is turned on, but plate voltage can be applied to the exciter only through the external relay terminals. In the third position, plate voltage is applied directly. In the fourth position the modulator is turned on and plate voltage applied through the relay terminals.

Construction

In assembling the exciter in a so-called amplifier-foundation enclosure with a 6 × 14-inch chassis, an effort has been made to keep the unit as compact as possible. If space is available, the constructor may wish to use a 7 × 17-inch enclosure which allows more space in which to work. To permit removal of the cover without disturbing the dial, the spot welds at one end of the cover are broken by drilling them out. Self-tapping screws are then substituted as fastenings. The dial (a National type SCN) is fastened to the cover end plate with the lower edge of the dial at chassis-top level. The side edges of the dial escutcheon plate are trimmed to fit the width of the cover, if necessary.

Looking at the rear view of Fig. 6-86, the condensers in the tuning gang, C_{10}, C_{22} and C_{27}, are mounted on a strip of polystyrene 3 inches wide and as long as necessary to accommodate the length of the gang. The polystyrene strip is supported on metal pillars at the corners to bring the gang shaft up level with the dial hub. The sockets for the r.f. tubes are lined up along one edge of the chassis with the two miniature modulator tubes along the opposite edge. The power transformer is

Fig. 6-85 — The gang-tuned bandswitching exciter is built in a standard amplifier-foundation enclosure. Additional holes have been punched in the top to aid ventilation.

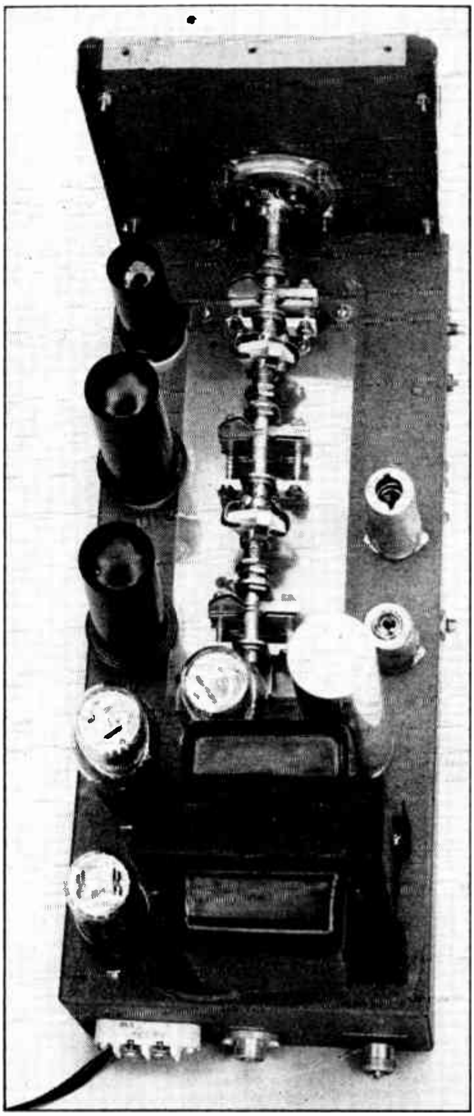


Fig. 6-86 — Rear view of the gang-tuned bandswitching exciter.

placed at the extreme rear end of the chassis and the filter input condenser, C_{32}, and the sockets for the rectifier and the two voltage-regulator tubes are grouped around in the remaining available space.

Underneath, the bandswitch, with the ceramic sections spaced out 2¾ inches, is mounted along the center line of the chassis. A metal bracket inside the last section holds the rear end of the assembly. The oscillator bandset condenser, C_9, is fastened to the side of the chassis, opposite the first bandswitch section where it can be adjusted with a screwdriver from outside. The deviation control, R_6, is similarly placed opposite the last bandswitch section.

The control switch and the two key jacks are to the right of the bandswitch control at the front end of the chassis. The two doubler bandset condensers, C_{23} and C_{28}, are on the right side of the chassis in Fig. 6-88, near the last bandswitch section. C_{23} is mounted vertically, with its shaft protruding out through the top of the chassis, while C_{28} is mounted horizontally so that it can be adjusted from the side. The two filter chokes occupy the rear of the chassis.

All coils are grouped around the bandswitch sections. Most of them can be supported by their leads from the switch. The oscillator coils are braced against vibration by cementing them, where necessary, to polystyrene-strip braces fastened to the chassis. An aluminum partition to the left of the bandswitch in Fig. 6-88 shields these coils from the others. It is advisable to cut or wind the coils with an extra turn or two for adjustment. The final trimming of inductance can be done by bending or sliding the last turn or two away from the other turns.

All power wiring should be done with shielded wire with the braid grounded as often as convenient. The only additional v.h.f. filtering found desirable in TVI tests was the installation of capacitors C_{30} and C_{31} across the a.c. line.

Adjustment

The tracking of the three stages is not difficult, but it should be done carefully. Only the last stage needs to be tracked for 3.5- and 7-Mc. operation, of course. A milliammeter should be placed temporarily in the circuit to read plate current to the 6L6s. The oscillator should first be adjusted with the bandswitch in the first (80-meter) position. Then, with C_{10} set at or close to maximum capacitance, C_9 should be adjusted until the oscillator is heard at 3500 kc. C_{10} is then turned to minimum capacitance and the frequency noted. If 4000 kc. comes too far inside the tuning range,

COIL TABLE FOR GANG-TUNED BAND-SWITCHING EXCITER

Coil	*Wire No.*	*Diam.*	*Length*	*Turns*
L_1	24	1″	1⅝″	50
L_2	24	⅝″	1 7/16″	47
L_3	20	1″	1 13/16″	29
L_4	20 enam.	⅝″	¾″	23
L_5	24	½″	½″	16½
L_6	24	½″	⅜″	11
L_7	22 enam.	1″	⅞″	32
L_8	20 enam.	⅝″	1 1/16″	21
L_9	24	½″	⅜″	11
L_{10}	20	¾″	9/16″	9
L_{11}	20	¾″	9/16″	4½

L_1 — B & W Miniductor No. 3016, L_2 — No. 3008, L_3 — No. 3015, L_5, L_6, L_9 — No. 3004, L_{10} — No. 3011, L_{11} — No. 3010.

Links as follows: L_7, L_8 — 5 turns, L_9 — 3 turns, L_{10}, L_{11} — 2 turns.

L_4 and L_7 close-wound.

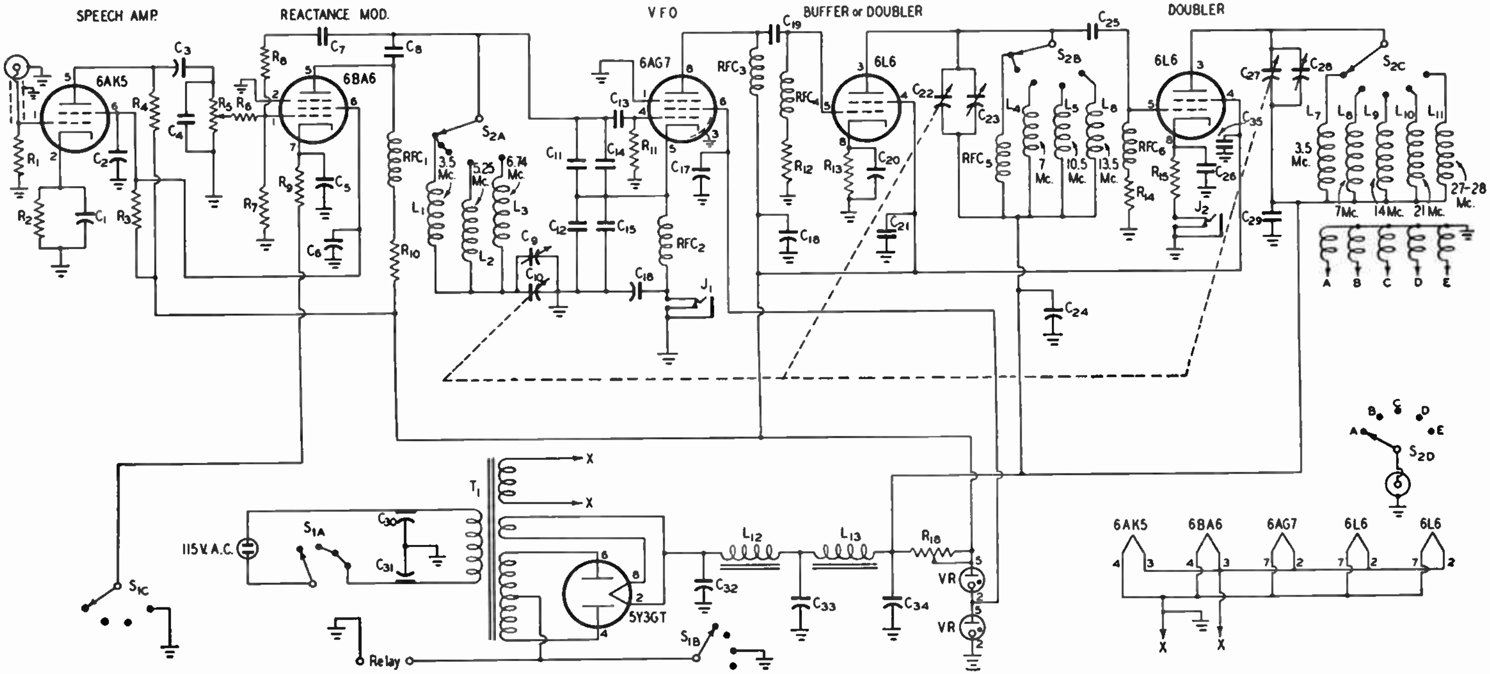


Fig. 6-87 — Circuit diagram of the gang-tuned bandswitching exciter.

C_1 — 25-μfd. 25-volt electrolytic.
C_2, C_3, C_5, C_{16}, C_{17}, C_{18}, C_{20}, C_{24} — 0.01-μfd. paper.
C_4 — 220-μμfd. mica.
C_6 — 8-μfd. 150-volt electrolytic.
C_7, C_8, C_{25} — 47-μμfd. mica.
C_9, C_{23}, C_{28} — 50-μμfd. air trimmer (Millen 26050).
C_{10}, C_{22}, C_{27} — 25-μμfd. variable (Millen 19025).
C_{11}, C_{12} — 680-μμfd. silvered mica.
C_{13}, C_{19} — 100-μμfd. mica.
C_{14} — 50-μμfd. 330 p.p.m. neg. coefficient condenser.
C_{15} — 25-μμfd. 330 p.p.m. neg. coefficient condenser.
C_{21}, C_{26}, C_{29}, C_{35} — 0.01-μfd. ceramic (Sprague 36C-1).
C_{30}, C_{31} — 0.01-μfd. paper (Sprague Hypass).
C_{32}, C_{33}, C_{34} — 8-μfd. 450-volt electrolytic.
R_1 — 2.2 megohms, ½ watt.
R_2 — 2200 ohms, ½ watt.
R_3, R_6 — 0.1 megohm, ½ watt.
R_4 — 0.22 megohm, ½ watt.
R_5 — ½-megohm potentiometer.
R_7 — 0.47 megohm, ½ watt.
R_8, R_{10} — 10,000 ohms, ½ watt.
R_9, R_{13}, R_{15} — 470 ohms, 1 watt.
R_{11} — 47,000 ohms, 1 watt.
R_{12} — 0.1 megohm, 1 watt.
R_{14} — 47,000 ohms, 1 watt.
R_{16} — 10,000 ohms, 10 watts.
L_1, L_2, L_3, L_4, L_5, L_6, L_7, L_8, L_9, L_{10}, L_{11} — See table.
L_{12}, L_{13} — 25-hy. 125-ma. filter choke.
J_1, J_2 — Closed-circuit jack.
RFC_1, RFC_3, RFC_4, RFC_6 — 2.5-mh. r.f. choke.
RFC_2 — 10-mh. r.f. choke.
RFC_5 — 10-μh. r.f. choke (National R-33).
S_1 — 3-pole 4-position rotary switch (Mallory 3134J).
S_2 — 3-section, 2 poles per section, 6-position ceramic rotary switch (Centralab Switchkit).
T_1 — Power transformer: 370-0-370 volts r.m.s., 100–150 ma.; 6.3 volts, 5 amp.; 5 volts, 3 amp. (Thordarson 22R33).
VR — VR-75-30 voltage-regulator tube.

L_1 should be decreased slightly and the process repeated, readjusting C_9 to bring the signal at 3500 kc. with C_{10} set at maximum. If, on the other hand, the tuning range is not sufficiently wide to include 4000 kc. when C_{10} is at minimum capacitance, L_1 must be increased and the above process repeated. The larger L_1 is, the greater will be the frequency range covered; the smaller L_1 is made, the greater the bandspread, i.e., the smaller the frequency range.

With the oscillator tuned to 3500 kc., C_{27} will be set at or near maximum capacitance, since it is ganged to C_{10}. C_{28} should now be adjusted to resonance, using the dip in 6L6 plate current as the indicator. Then with the oscillator tuned to 4000 kc., resonance in the output circuit should again be checked. If the tuning is not at resonance, it should be carefully observed whether C_{28} must be increased or decreased to restore resonance. If C_{28} must be decreased, C_{27} is not tuning fast enough. In this case, the size of L_7 must be increased slightly and the process repeated. If an increase, instead of a decrease, in the capacitance of C_{28} is required to restore resonance, L_7 must be trimmed down.

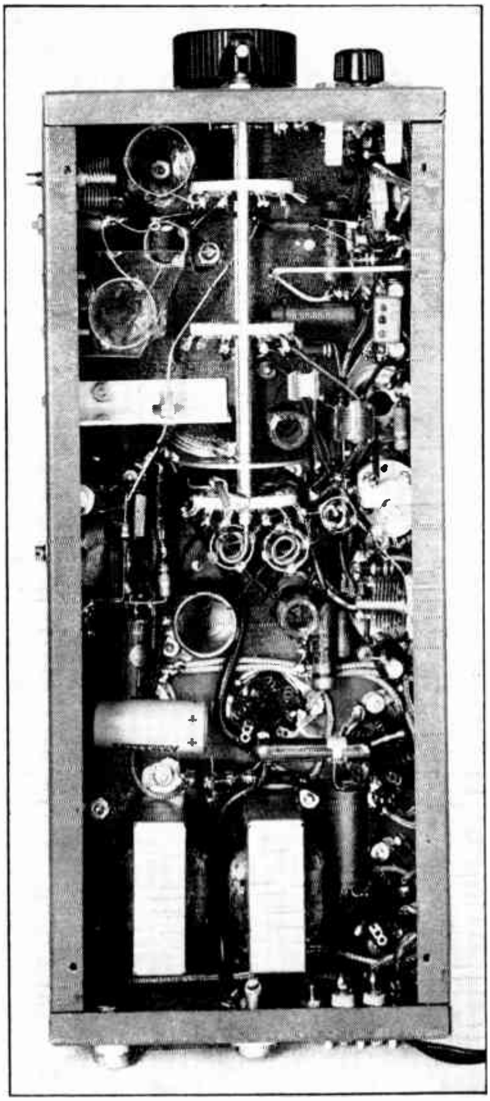


Fig. 6-88 — Bottom view of the gang-tuned bandswitching exciter. The chassis is 3 inches deep.

With L_1 and L_7 adjusted so that the oscillator and output stages track over the 80-meter band, the settings of C_9 and C_{28} should be noted carefully so that they may be reset at the same points. L_8 is next adjusted so that the output circuit, with the bandswitch in the 40-meter position, tunes to 7000 kc. at the same setting of the gang that tunes to 3500 kc. C_{28} should not be readjusted for this except experimentally to determine if L_8 need be increased or decreased to tune the circuit to 7000 kc. When the adjustment is completed, all condenser settings should be the same for 7000 and 3500 kc. If this is done, the circuits should track over the 40-meter band without further adjustment.

With the switch in the 20-meter position, L_4 is adjusted next. With the gang set so that the oscillator tunes to 3500 kc., C_{23} should be adjusted for resonance at 7000 kc. Then, with the oscillator tuned to 3600 kc., C_{23} should be adjusted for resonance at 7200 kc. If C_{23} must be changed from its original setting to restore resonance, L_4 must be adjusted following the procedure outlined previously for the adjustment of L_7.

Next, L_9 is adjusted so that the output circuit tunes to 14,000 kc. when the oscillator is set at 3500 kc., *without disturbing the setting of C_{28}*. The output circuit should then track at least up to 14,400 kc., the highest frequency of interest.

With the switch in the 21-Mc. position, L_2 is connected into the oscillator circuit. This coil should be adjusted so that the oscillator signal is heard at 5250 kc. when the gang is set near maximum capacitance. *The setting of C_9 should not be disturbed.* Then, without disturbing the setting of the gang or of C_{23}, L_5 should be adjusted so that the circuit resonates at 10,500 kc. and L_{10} should be adjusted so that the output circuit tunes to 21,000 kc. without disturbing the setting of C_{28}. The three stages should then track over the full width of the 21-Mc. band.

The adjustment for 27–28 Mc. is similar. L_3 is trimmed so that the oscillator tunes to 6740 kc. with the gang near maximum capacitance. At the same setting, L_6 is adjusted for 13,480 kc. and L_{11} for 26,960 kc.

The power supply shown delivers a voltage of 360 under load. At this voltage, the 6AG7 cathode current should run 10 ma. or less, while the cathode current in the first and second doubler stages should be approximately 25 and 35 ma. respectively.

A 175-Watt Transmitter for the 160-Meter Band

A single transmitter that will cover the extremes of 1.8 and 28 Mc. necessarily must involve considerable compromise as well as complication. From several considerations, it is not only preferable, but also economically feasible, to build a separate unit for 160 meters, since it can be simple and straightforward. In most instances, operating conditions may be chosen so that the 160-meter unit will operate from the same power supply as the higher-frequency transmitter, if the station has one.

An example is shown in Figs. 6-89 through 6-93. Because the 1.8-Mc. band is divided into narrow slices, crystal control is preferable to reduce the danger of out-of-band operation. The oscillator circuit in this case is a modified Pierce with a separate untuned plate output circuit. C_3 is a feed-back-adjustment condenser.

The 6L6 stage provides the necessary buffering between the oscillator and the final amplifier for 'phone operation. There is no danger of oscillation at the fundamental in the buffer stage because its input circuit is untuned. Since the frequency range to be covered is small, the output circuit of this stage is easily broadbanded. Thus only a single tuning control is required for the entire transmitter.

The triode final amplifier is a conventional arrangement with a capacitive-divider plate neutralizing circuit. The d.c. connection to the rotor of the tank condenser through RFC_6 makes it possible to use a condenser with half the peak-voltage rating that otherwise would be required. RFC_5 is a v.h.f. parasitic suppressor.

For c.w. operation, the oscillator is keyed in the cathode circuit. R_5 provides protective bias for the buffer stage.

The layout is suitable for any of the usual triodes with plate-cap connection, operating at plate voltages up to 1500 with a plate-voltage/plate-current ratio of 10 or greater. If a tube with a 6.3-volt filament is chosen, only a single filament transformer is needed.

Construction

The unit is assembled on an 8 × 17 × 3-inch chassis with an 8¾-inch panel. Most of the constructional details are evident from the photographs. The output-stage tank condenser is mounted on ceramic pillars and its shaft is fitted with an insulating coupling. The condenser is placed on the chassis so that its dial and the milliammeter will be symmetrical in respect to the center of the panel. The tank coil is a homemade affair wound in two equal sections on separate Millen type 44000 polystyrene forms, each cut down to a length of 2½ inches. The outer end only of each section is fastened to a 1¼-inch cone insulator, and the two sections are placed with their inner ends an inch apart. Additional bracing is provided by the No. 14 wire leads from the inner end of each section to the plate r.f. choke, RFC_6, mounted near the center. After winding the turns are cemented in place with coil dope.

The output link (8 turns of No. 18 d.c.c. should be satisfactory) is wound on a ¾-inch length of leftover form. A length of ¼-inch polystyrene rod is cemented to the inside surface of the link form. This shaft then runs through a panel bearing fitted with a National type RSL shaft lock which provides an adjustable friction for the shaft. A knob on the shaft provides a means of adjusting the coupling from the panel.

The neutralizing condenser, C_N, is placed close to the tube, between the tube and the panel. C_{15} should not be less than the value specified, nor larger than 0.005 μfd., if the amplifier is to be plate-modulated.

All components for the exciter stages, except the two tubes and crystal, are placed underneath the chassis. These include the plate tank circuit of the buffer stage. C_7 is mounted so that it may be adjusted with a screwdriver from on top. L_1 is wound on a Millen 1-inch plastic form and is placed alongside the con-

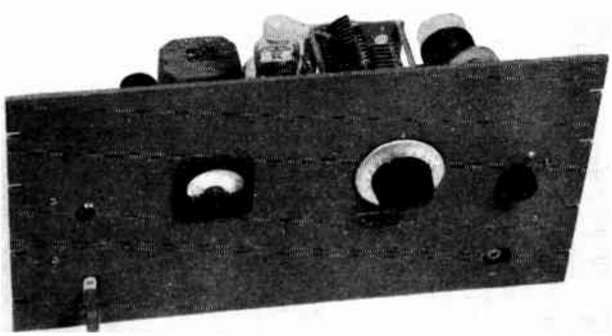


Fig. 6-89 — **Front view of a 175-watt transmitter for the 160-meter band. Only one tuning control is needed, plus a small knob used to adjust the setting of the swinging link on the output coil.**

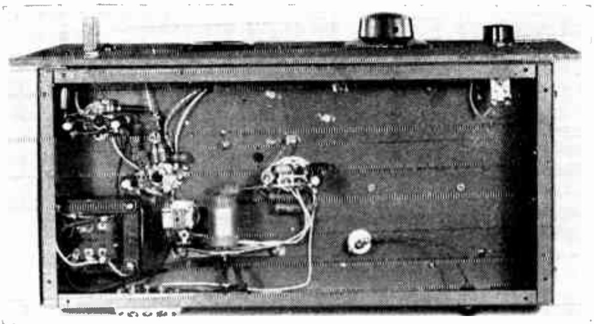


Fig. 6-90 — Bottom view of the 160-meter transmitter. The oscillator tube socket and its related parts are in the upper left corner. The 6L6 and the final-amplifier tube are mounted in a line through the center of the chassis, with the plate coil for the 6L6 supported on a bracket between the two stages. The parasitic-suppressing choke is mounted between the grid terminal of the amplifier socket and a ceramic stand-off insulator.

denser on a bracket that spaces it from the chassis on all sides.

For convenience in changing crystals, the crystal socket is mounted on the front edge of the chassis, at the left. Clearance holes for both the crystal socket and the key jack are cut in the panel.

The placement of the filament transformers is not critical, except that they should not be so far from the tube sockets that excessive voltage drop results through the wiring. In this case it was convenient to place one on top of the chassis and the other below. Terminals are provided across the back for high voltage, low voltage, bias and ground. An a.c. cord makes the line-voltage connection to the filament-transformer primaries.

Fig. 6-92 shows the diagram of a suitable power supply in case a separate supply is necessary or desirable. Control circuits are included.

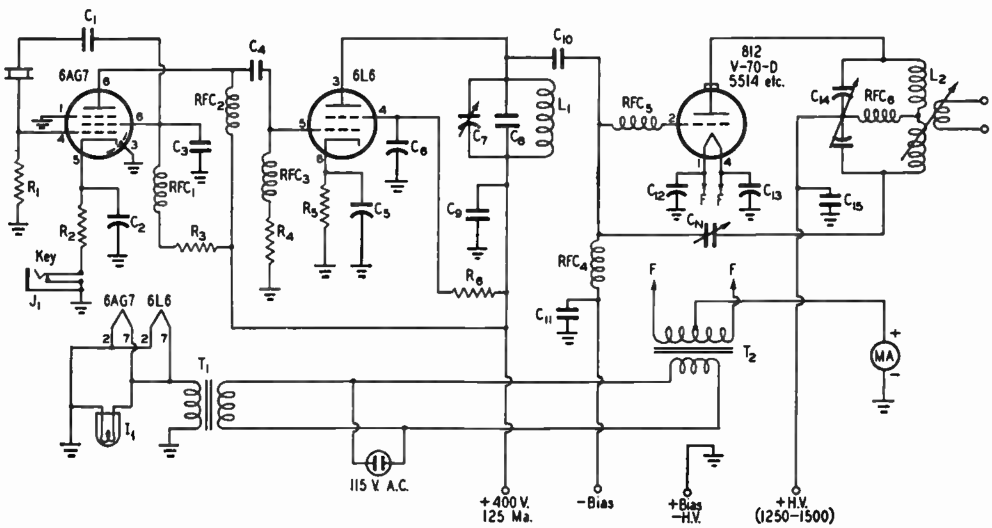


Fig. 6-91 — Schematic diagram of a single-control 175-watt transmitter for the 160-meter band.

C_1 — 0.001-μfd. mica, 400 volts.
C_2, C_5, C_6, C_{12}, C_{13} — 0.01-μfd. 600-volt paper.
C_3 — 10-μμfd. mica. See text.
C_4, C_8 — 100-μμfd. mica.
C_7 — 50-μμfd. variable (National PSR-50).
C_9, C_{11} — 0.0068-μfd. mica, 500 volts.
C_{10} — 220-μμfd. mica, 600 volts.
C_{14} — 100-μμfd.-per-section dual transmitting variable, 0.070 air gap (3000 volts peak). (National TMC-100-D.)
C_{15} — 0.0035-μfd. mica, 5000 volts.
C_N — Neutralizing condenser, 0.8–10 μμfd. (NC-800-A).
R_1 — 15,000 ohms, ½ watt.
R_2 — 330 ohms, 1 watt.
R_3 — 39,000 ohms, 1 watt.
R_4 — 22,000 ohms, ½ watt.
R_5 — 600 ohms, 2 watts (two 1200-ohm 1-watt units in parallel).
R_6 — 10,000 ohms, 5 watts.
L_1 — 46 turns No. 26 d.s.c. close-wound on 1-inch diam. form.
L_2 — Each half consists of 46 turns No. 20 d.s.c. close-wound on a 1⅞-inch diam. form (Millen 44000). The two halves are mounted so that there is 1⅛ inches between windings to permit passage of the link coil. Link: 8 turns No. 18 d.c.c. close-wound on 1⅞-inch diam. form made of same material as the main coil form.
I_1 — 6.3-volt panel lamp.
J_1 — Closed-circuit jack.
MA — 0–300 ma. d.c. meter.
RFC_1 through RFC_4 — 2.5-mh. r.f. choke (National R-100-S).
RFC_5 — 21 turns No. 26 d.s.c. close-wound on ¼-inch diam. form (a 1-watt resistor of any high value may be used as the form).
RFC_6 — Transmitting r.f. choke (Millen 34140).
T_1 — 6.3-volt 3-amp. filament transformer (Stancor P-6014).
T_2 — 7.5-volt 4-amp. filament transformer (UTC S-56).

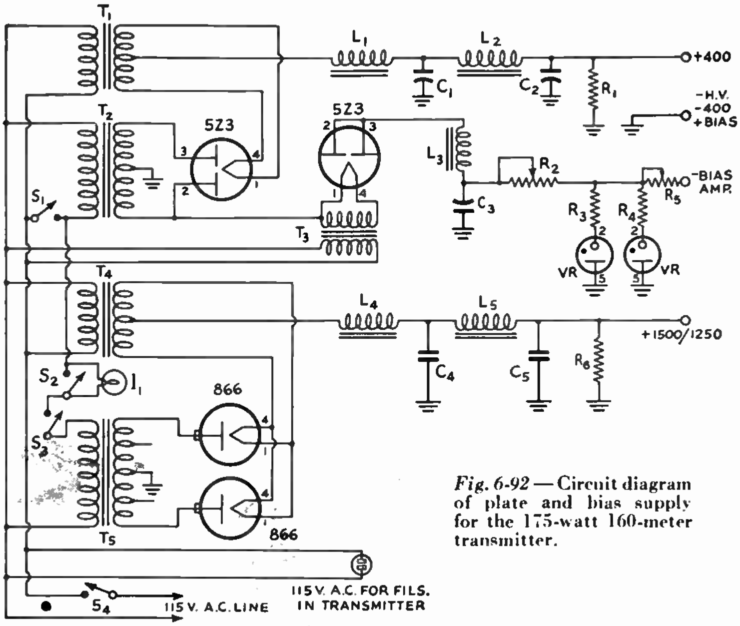


Fig. 6-92 — Circuit diagram of plate and bias supply for the 175-watt 160-meter transmitter.

C_1, C_2, C_3 — 8-μfd. 600-volt-wkg. elec.
C_4, C_5 — 4-μfd. 2000-volt oil-filled.
R_1 — 25,000 ohms, 25 watts.
R_2 — 30,000 ohms, 10 watts, with slider.
R_3, R_4 — 47 ohms, 1 watt.
R_5 — See text.
R_6 — 25,000 ohms, 150 watts.
L_1 — 5/25-hy. 150-ma. swinging choke.
L_2 — 20-hy. 150-ma. smoothing choke.
L_3 — 30-hy. 75-ma. filter choke.
L_4 — 5/25-hy. 200-ma. swinging choke.
L_5 — 10-hy. 200-ma. smoothing choke.
I_1 — 150-watt 115-volt lamp.
S_1, S_3, S_4 — 10-amp. toggle switch.
S_2 — 5-amp. toggle switch.
T_1, T_3 — 5-volt 3-amp. filament transformer.
T_2 — 400-v. d.c. 225-ma. plate transformer.
T_4 — 2.5-volt 10-amp. filament transformer, 10,000-volt insulation.
T_5 — 1750/1500/1250-v. d.c. 200-ma.-or-more plate trans.
VR — VR-75 voltage-regulator tube.

Adjustment

If the transmitter is to be used for c.w. operation, it may be desirable to experiment briefly with C_3 to obtain best keying characteristics. It may be found that a different capacitance will work better with some crystals, while with others the condenser may not be needed at all, or that the keying will be better with C_3 connected from grid to ground, rather than from screen to ground.

With the oscillator running, the d.c. voltage across the buffer grid leak, R_4, should be 90 to 110 volts. A milliammeter placed in the 400-volt lead will read the combined currents of the oscillator and buffer (100 to 120 ma.). However, there should be a usable dip in current when C_7 tunes the buffer tank circuit through resonance. If the circuit is tuned to 1850 kc., it will not need readjustment for any frequency between 1800 and 1900 kc. Similarly, if it is initially adjusted for 1950 kc., it will cover the 1900- to 2000-kc. range.

The proper bias adjustment for the final amplifier will depend upon the type of tube used. Any additional operating bias voltage above 75 volts is obtained by grid-leak action from R_5 in the power supply. The resistance at which R_5 should be set can be determined by subtracting 75 from the rated operating bias for the tube used and dividing the remainder by the rated grid current in *amperes*. The amplifier should be neutralized before applying plate voltage. If necessary, the size of L_2 should be adjusted so that resonance occurs with the tank condenser set near maximum capacitance.

The choke, RFC_5, should be the only means necessary to suppress v.h.f. parasitic oscillation if a Type 5514 tube is used. Other tubes may require circuit alterations.

Fig. 6-93 — Rear view of the 1.8-Mc. transmitter. The construction of the amplifier plate coil and its swinging link is shown at the left. The plate r.f. choke and the plate by-pass condenser are mounted underneath the main tuning condenser, which is supported by 1-inch stand-off insulators. An insulated coupling is used between the rotor shaft of the condenser and the panel control. The neutralizing condenser is visible behind the amplifier tube.

A Push-Pull 807 Amplifier with Multiple-Band Tuners

A push-pull 807 amplifier and antenna tuner requiring no plug-in coils or bandswitching is shown in Figs. 6-94 through 6-99. The tanks of both the amplifier and antenna tuner are made of the new multiband circuits — combination circuits that show multiple resonances through the range of the tuning condensers. All bands, 80 through 10 meters, are covered as the ganged condensers are turned through their capacitance range. Tuners of this type are available on the market, or they can be built.

Referring to the diagram of Fig. 6-96, the grid-tank tuner is made up of C_1, C_2 and L_1, L_2 and L_3. C_3, C_4, L_6, L_7 and L_8 comprise the plate-tank tuner. The antenna tuner consists of C_5, C_6, L_{11}, L_{12} and L_{13}. The output tank circuit is coupled to the antenna tuner both through a link line for the high frequencies and by the coupling coils L_9 and L_{10}, which are included in the manufactured units, for low frequencies. Series or parallel tuning may be used for 3.5 and 7 Mc. Parallel tuning is used for the higher bands.

RFC_1 and RFC_2 are parasitic suppressors. C_{10} and C_{11} are tubular air condensers connected directly between plate and cathode. They contribute to v.h.f. harmonic reduction as well as parasitic suppression. L_4C_7 and L_5C_8 are v.h.f. traps that may be tuned to the second or third harmonic of 28-Mc. band frequencies as found necessary to reduce TVI. The 6Y6G reduces the power input to the 807s to a safe value when excitation is removed.

Construction

The amplifier is constructed on a 10 × 12 × 3-inch chassis with a 11 × 12 × 8-inch utility box as an inexpensive shielding cabinet. The box, with its bottom cover removed, is fastened to the chassis by means of 1-inch metal-strip cleats along each 10-inch side of the chassis. The box then overhangs the chassis at the rear, providing, together with the holes in the top cover, ventilation essential in an enclosed job such as this. The spare bottom cover is cut to form a partition shielding the amplifier plate tank from the antenna tuner. A door cut in the back provides access to the antenna-tuner terminals.

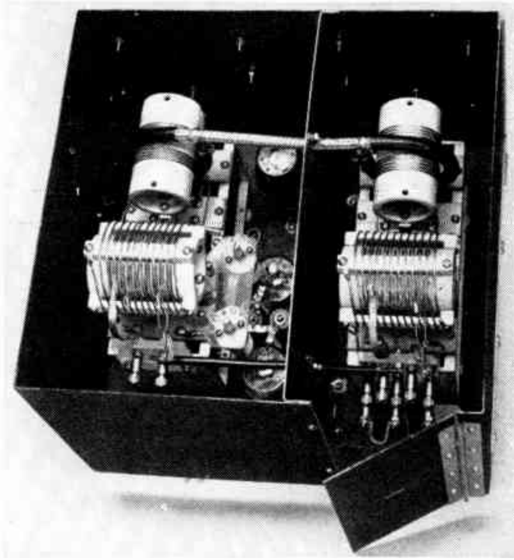


Fig. 6-95 — Rear-top view of the multiband 807 amplifier, showing the hinged door in the rear for altering antenna connections. The amplifier is to the left, the antenna tuner to the right. The third tube is the 6Y6G. The outer conductor of both link lines is grounded on the shielding partition.

The two large tuners are mounted on 1-inch cone insulators. The harmonic traps, C_7L_4 and C_8L_5, are mounted on a strip of polystyrene fastened to the plate tuner with brackets. Clearance holes for the 807s are cut in the top of the chassis alongside the plate tuner. The tube sockets are submounted in an aluminum strip 2½ inches wide spanning the chassis. The edges of this strip

Fig. 6-94 — A multiband push-pull 807 amplifier and antenna tuner with no plug-in coils or switching. The aluminum strip on top covers the harmonic-trap adjusting holes. The other holes are for ventilation. The panel is a sheet of 1/16-inch aluminum 12¼ inches wide and 10½ inches high.

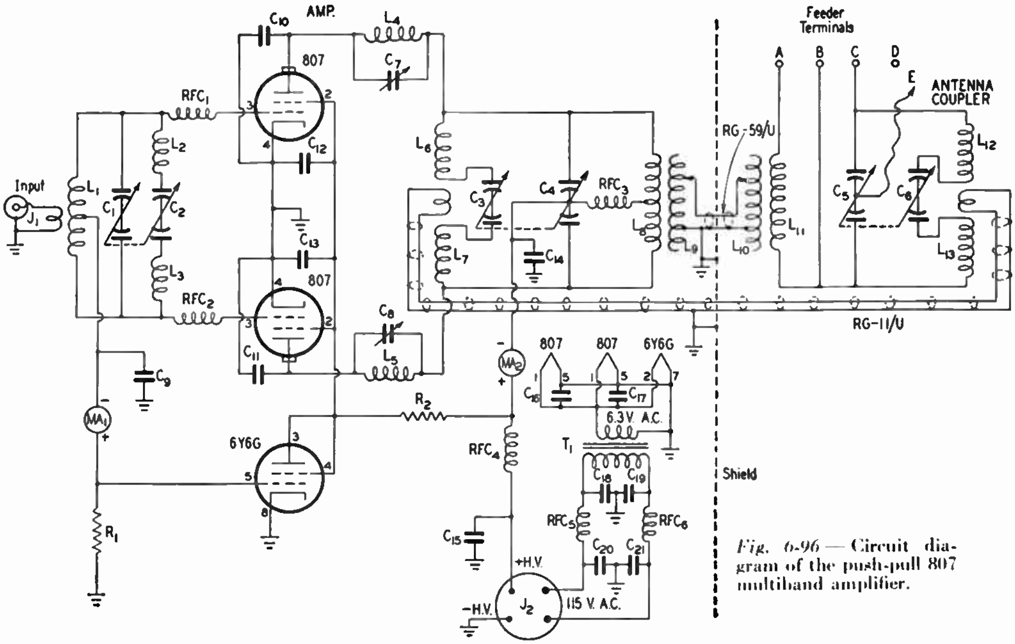


Fig. 6-96 — Circuit diagram of the push-pull 807 multiband amplifier.

C_1, C_2 — 125-μμfd. variable (National SSH-125), part of MB-50 tuner).
C_3, C_4, C_5, C_6 — 110-μμfd.-per-section variable (part of National MB-150 tuner).
C_7, C_8 — 50-μμfd. variable (National PSE).
C_9 — 0.0047-μfd. mica.
C_{10}, C_{11} — 10-μμfd. tubular; see text.
C_{12}, C_{13} — 0.005-μfd. ceramic (Centralab DA-048).
C_{14} — 0.001-μfd. mica, 1200 volts working.
C_{15} — 500-μμfd. mica, 1200 volts working.
C_{16}, C_{17}, C_{18}, C_{19}, C_{20}, C_{21} — 470-μμfd. mica.
R_1 — 12,000 ohms, 1 watt.
R_2 — 25,000 ohms, 20 watts.
L_1 — 30 turns No. 22 enam., center-tapped, 1¼ inches long, 1-inch diam.
L_2, L_3 — 7 turns No. 22 enam., 5⁄16 inch long, 1-inch diam., with ⅜-inch space between sections.
(NOTE: Above coils are part of MB-50 tuner.)
L_4, L_5 — 4 turns No. 16 tinned, 1 inch long, 5⁄16-inch diam.
L_6, L_7, L_{12}, L_{13} — 5 turns No. 12, ⅝ inch long, 1¾-inch diam., with ⅜-inch space between sections.
L_8, L_{11} — 18 turns No. 12, 2 inches long, 1¾-inch diam.; L_8 is center-tapped.
L_9, L_{10} — 12 turns No. 12, 2½ inches long, 2½-inch diam. (NOTE: L_6 to L_{13}, inc. — part of MB-150 tuner.)
J_1 — Coaxial-cable connector.
J_2 — 4-prong male plug.
MA_1 — 0–25 d.c. milliammeter.
MA_2 — 0–300 d.c. milliammeter.
RFC_1, RFC_2 — 1-μh. r.f. choke (National R33).
RFC_3 — 2.5-mh. r.f. choke (part of MB-150 tuner).
RFC_4, RFC_5, RFC_6 — 7-μh. r.f. choke (Ohmite Z50).
T_1 — 6.3-volt 3-amp. filament transformer (Stancor P5014).

are bent up a quarter of an inch to provide longitudinal strength.

Underneath, the grid tuner is mounted close to the tube sockets on pillars that space the coils evenly between the top and bottom of the chassis. The shaft of this tuner is operated by a pulley system so that the control can be brought out to the center of the panel. The pulleys are easily made by lightly grooving small bakelite tuning knobs.

The tubular condensers, C_{10} and C_{11}, are made as shown in the sketch of Fig. 6-97. They are mounted between the two tube sockets after clearance holes have been cut in the aluminum strip and the top of the chassis. The harmonic-filter components are placed close to the associated power terminals at the back. All power wiring should be done using shielded wire.

Modification of Antenna Tuner

If the National MB-150 tuner is used, slight modification is necessary to adapt it to the antenna tuner. The r.f. choke is removed. One end of L_{11} is disconnected from the condenser and is brought to one of the antenna terminals as indicated in Fig. 6-96. The original coupling clips are removed from L_9 and L_{10}. On each of the two coupling coils, one of the flexible leads is soldered permanently to the third turn from one end. The other lead is terminated in a copper spring clip. The original clips are fastened permanently to the coil after the position of the taps for proper coupling has been determined. These then serve as points or taps to which the spring clip can

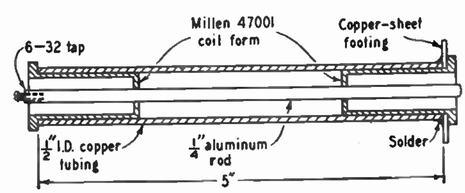


Fig. 6-97 — Sketch showing the construction of the tubular air condensers used in the p.p. 807 amplifier. The condenser is mounted with screws through the footing.

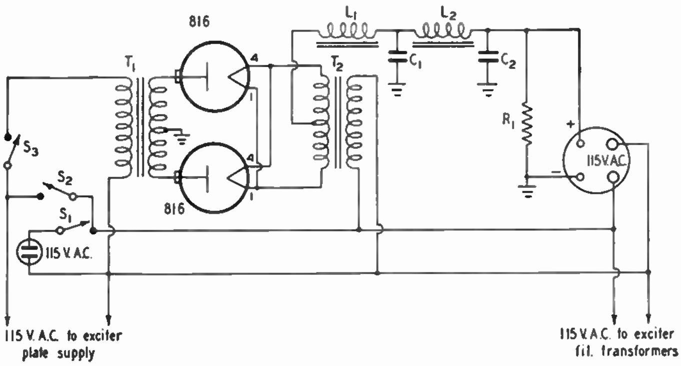


Fig. 6-98 — Circuit of power supply for the push-pull 807 amplifier.
C_1, C_2 — 4-μfd. 1000-volt oil-filled.
R_1 — 25,000 ohms, 50 watts.
L_1 — 5/25-hy. 250-ma. swinging choke.
L_2 — 10-hy. 250-ma. smoothing choke.
S_1 (filaments), S_2 (low voltage), S_3 (high voltage) — S.p.s.t. toggle switch.
T_1 — Plate transformer: 600/750 volts d.c., 250 ma.
T_2 — Filament transformer: 2.5 volts, 1 amp.

be quickly attached in changing bands. RG59/U cable is satisfactory for the low-frequency coupling line between L_9 and L_{10}, but the larger RG11/U should be used for the high-frequency line to avoid breakdown. Each end of the latter is formed into a coaxial link (see "Stray Coupling," Chapter Ten).

Adjustment

The diagram of a power supply for this amplifier is shown in Fig. 6-98. Approximately three watts of driving power is required.

If the National units are used, or if the tuners have been built closely to the specifications given under Fig. 6-96, adjustment of the amplifier should be merely a matter of resonating the input circuit for maximum grid current for the desired band, and the plate circuit for minimum plate current in the usual fashion (see "Adjustment of R.F. Amplifiers," this chapter). However, as the tank condensers are turned from minimum to maximum capacitance, the tuners do not resonate in the bands in logical sequence, but as follows: 28–27 Mc., 7 Mc., 21 Mc., 3.5 Mc. and 14 Mc. Minor resonances may be found at other multiples of frequencies that may be fed through from a multiband exciter. Therefore, until the dials have been plainly marked, the frequency should always be checked with a wavemeter.

The excitation should be adjusted for a grid current of 8 ma. under load. This should give a bias of 90 volts under full-load operating conditions. Without load, the plate current should dip to about 20 ma. With load, the coupling may be increased until the plate current at resonance is 200 ma. when the screen current should be approximately 16 ma. at 300 volts. If excitation is removed, the 6Y6G should limit the plate current to 35 ma. or less.

ANTENNA-COUPLER CONNECTION CHART FOR THE 807 P.P. AMPLIFIER

Tuning	*"C"*	*Feeder Terminals*	*Jump Terminals*	*Connect Clip Lead "E" to*
Series	Low	A & C	—	D
Series	Medium	A & C	—	B or C
Series	High	A & D	B & C	D
Parallel	Low	A & B	A & C	D
Parallel	Medium	A & B	A & C	B or C
Parallel	High	A & B	A & C	A

The antenna tuner is adjusted following conventional procedure (see "Practical Coupling Systems," Chapter Ten). The accompanying table shows how the terminal connections should be made to obtain the desired combination. If a center-fed antenna system is used whose total length for each half is a half wavelength for 3.5 Mc., or a multiple of that length, it should be possible to use parallel tuning on all bands, dispensing with the change in feeder connections to the antenna coupler.

♦

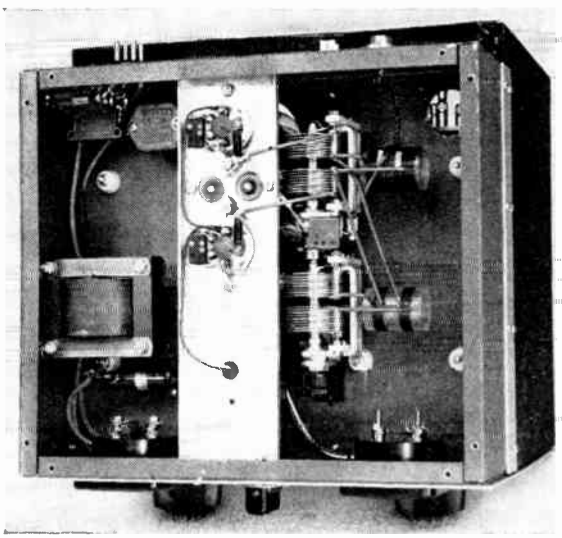


Fig. 6-99 — Bottom view of the push-pull 807 amplifier. The grid-circuit multiband tuner is to the right of the aluminum strip on which the 807 sockets and tubular condensers are mounted. The filament transformer is to the left with the v.h.f. power-lead filters above.

Two Triode Amplifiers

Figs. 6-100 through 6-110 show the circuits and photographs of two neutralized-triode amplifiers designed especially toward eliminating parasitic oscillation and the minimizing of v.h.f. harmonic radiation. The chief objective in the arrangement of components in each case is that of keeping the inductance of r.f. connections to the tubes as short as possible.

The design of both amplifiers as shown is suitable for any conventional plate-cap-connection triodes operating with a plate-voltage/plate-current ratio of 10 to 1 or greater per tube, at plate voltages up to 1500 volts.

The circuit diagram of the single-tube amplifier is shown in Fig. 6-102. It is conventional, except for the addition of the condensers C_5, C_9 and C_{10}, the v.h.f. wavetrap C_8L_3 in the plate lead, and the v.h.f. filters in the power leads. The trap may be adjusted to either the second or third harmonic of 28-Mc. band frequencies. C_5 is a small ceramic condenser connected directly between the grid and filament terminals of the tube socket for the purpose of providing a short low-inductance path across the input of the tube. C_9 serves the same purpose at the output of the tube, while C_{10} helps to maintain the original circuit balance as well as to provide a

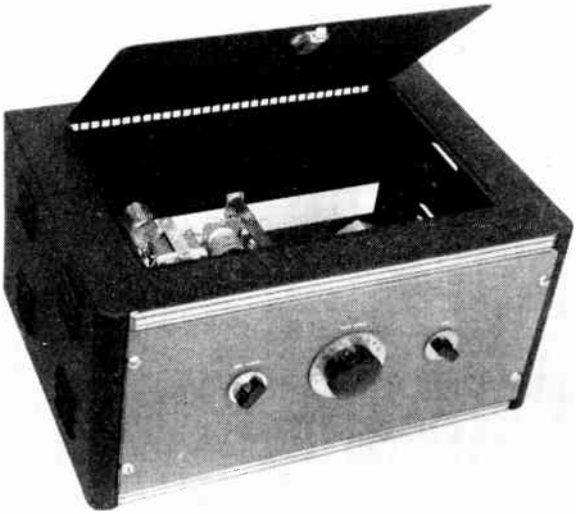


Fig. 6-100 — The single-tube triode amplifier installed in a metal cabinet. The controls are for input- and output-circuit tuning and variable-link adjustment. The panel is of metal 8¾ inches high.

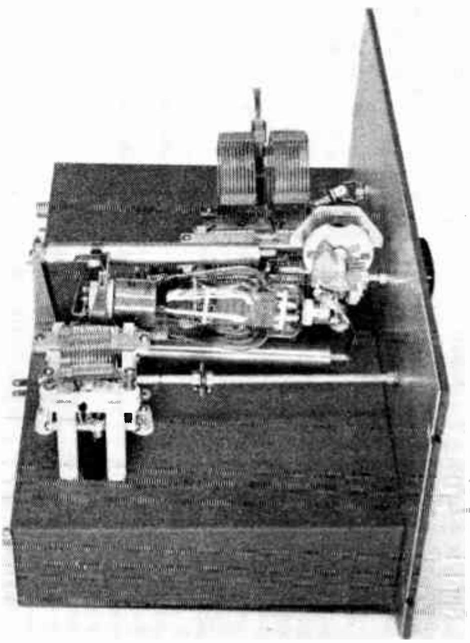


Fig. 6-101 — Grid-circuit end of the single-tube triode amplifier. The grid tank coil in the foreground is mounted on 2½-inch ceramic pillars to bring its terminals close to the terminals of the grid tank condenser immediately behind the coil. The two tubular condensers are visible, one below the tube and the other above and behind the tube. The air trimmer condenser close to the plate cap of the tube is the harmonic-trap tuning condenser. The panel in this case is a Presdwood section backed with aluminum sheet.

low-inductance lead to the neutralizing condenser. These two condensers are of the tubular air type.

The plate tuning condenser, C_2, is mounted on ½-inch cone insulators at the center of the 10 × 17 × 3-inch chassis, with just enough space at the front end for an insulating shaft coupling. A narrow aluminum bracket is made for the socket so that the tube will be held horizontally with its plate cap close to the front stator terminal of the tuning condenser.

The two tubular condensers are made similar to the sketch of Fig. 6-104 except that a ½-inch cone insulator is cemented in the tubing at one end in place of the button insulator. The ⅛-inch rod threads into the top of the cone. C_9 is mounted from the bracket supporting the tube socket so that it is orientated

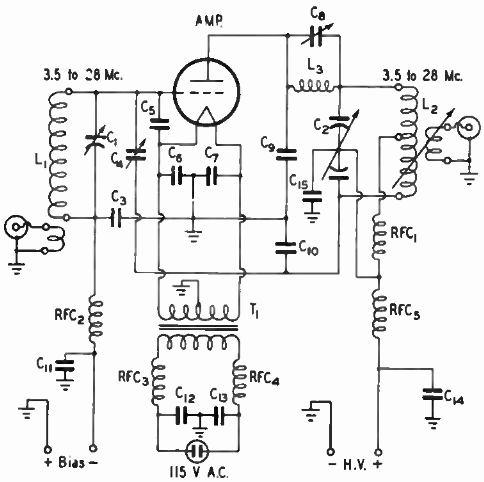


Fig. 6-102 — Circuit diagram of the single-tube triode amplifier.

C_1 — 100-μμfd. variable, 0.03- to 0.047-inch plate spacing (National TMK-100).
C_2 — Dual variable — 100-μμfd. per section, 0.017-inch spacing (National TMK-100-D).
C_3, C_6, C_7 — 0.001-μfd. mica.
C_4 — Disk-type neutralizing condenser (Hammarlund NZ-10).
C_5 — 12-μμfd. ceramic.
C_8 — 100-μμfd. variable air padder (Millen 17100).
C_9, C_{10} — Tubular fixed air condenser (see text and Fig. 6-104 for constructional details).
C_{11}, C_{12}, C_{13} — 470-μμfd. mica.
C_{14} — 500-μμfd. 5000-volt mica.
C_{15} — 0.001-μfd. 5000-volt mica.
L_1, L_2 — See coil table.
L_3 — 3 turns No. 16 tinned, ⅜-inch diam., ½ inch long.
RFC_1 — 1-mh. 300-ma. r.f. choke.
RFC_2, RFC_3, RFC_4, RFC_5 — 7-μh. r.f. choke (Ohmite Z-50).
T_1 — Filament transformer: 6.3 volts, 4 amp.

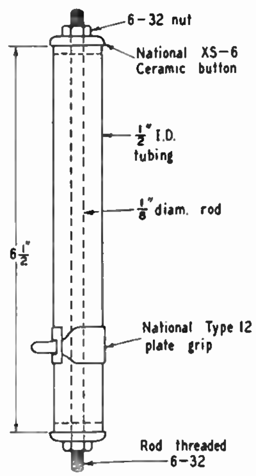


Fig. 6-104 — Sketch showing the construction of the tubular fixed condensers. The plate grip makes a convenient mounting and connection point.

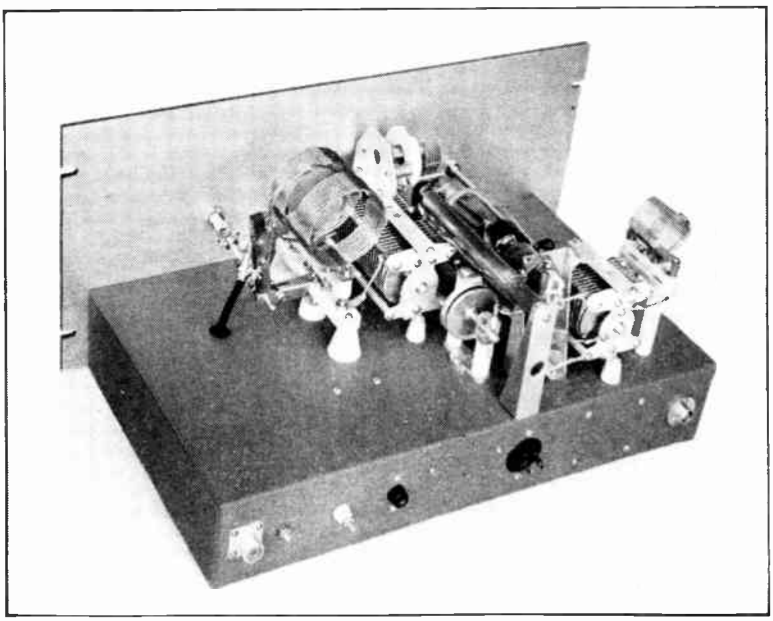


Fig. 6-103 — Rear view of the single-tube triode amplifier. The tubular condenser in view is C_{10}. The chassis is 10 inches deep.

COIL TABLE FOR SINGLE-TUBE TRIODE AMPLIFIER

Band	*Turns*	*Wire No.*	*Diam.*	*Length*	*Lμh.*
** Plate*					
27–28 Mc.	6	8	1¾″	2⅛″	1.2
21 Mc.	8	8	2″	2¾″	1.8
14 Mc.	10	14	2½″	2½″	4.0
7 Mc.	18	14	2½″	2¾″	12.0
3.5 Mc.	30	16	2½″	2¾″	36.0
*** Grid*					
27–28 Mc.	7	16	1¼″	1⅛″	1.2
21 Mc.	9	16	1¼″	¾″	2.2
14 Mc.	14	16	1¼″	1¼″	4
7 Mc.	22	18	1¼″	1¼″	10
3.5 Mc.	28	18	1¼″	⅞″	22

* B&W BVL series coils, unaltered except for 1 turn off each end of BVL-10. Halves of coils are separated ½ inch.

** National AR-16-E series coils with turns removed as follows: 28 Mc. — 1 turn, 14 Mc. — unaltered, 7 Mc. — 6 turns, 3.5 Mc. — 28 turns. AR-16-20E with 5 turns removed for 21 Mc.

directly beneath the tube and parallel to it.

A second aluminum bracket at the rear of the chassis supports one end of C_{10} so that the other end is close to the inside terminal of the rear stator section of C_2. A clamp around the outer conductor makes a connection to the bracket supporting the tube and provides additional bracing. The two filament by-pass condensers are grounded on the socket-mounting screws.

The neutralizing condenser, C_4, is tucked in between the bracket that holds C_{10} and the rear of the tuning condenser. It may be adjusted through a ½-inch hole cut in the bracket. The trap condenser, C_8, is mounted on a bracket fastened to the frame of the plate tuning condenser so that the stator terminal is close to the plate cap of the tube. A short lead connects the rotor terminal of C_8 to the front stator terminal of C_2. The trap coil, L_3, is soldered across C_8, underneath.

The grid tuning condenser is mounted on ½-inch cone insulators close to the tube socket. The grid-circuit by-pass condenser, C_3, is connected between the rear end plate of C_1 and the ground point at the tube socket.

On the other side, the plate tank-coil mounting is placed close to the tuning condenser. The plate r.f. choke, RFC_1, is mounted between the coil and condenser, and the high-voltage lead comes up from under the chassis through a ceramic bushing.

A control for the variable link is brought out to the panel to balance the grid tuning control on the other side. A pair of Millen universal-joint shaft couplings is used to take care of the displacement between the link shaft and the panel control.

The filament transformer and the harmonic-filter components are placed underneath the chassis close to the power terminals at the rear. All power wiring is done with shielded wire.

The Push-Pull Amplifier

The circuit of the push-pull amplifier is shown in Fig. 6-106. The same general principles are followed both in the circuit and in construction. Provision is made for metering the individual grid and cathode currents so that the amplifier may be adjusted for balanced current readings. In this amplifier tubular neutralizing condensers, made as shown in the photograph of Fig. 6-107, are used.

The plate tank condenser is placed on the chassis with its shaft 3½ inches from the end of the 10 × 17 × 3-inch chassis. It must be

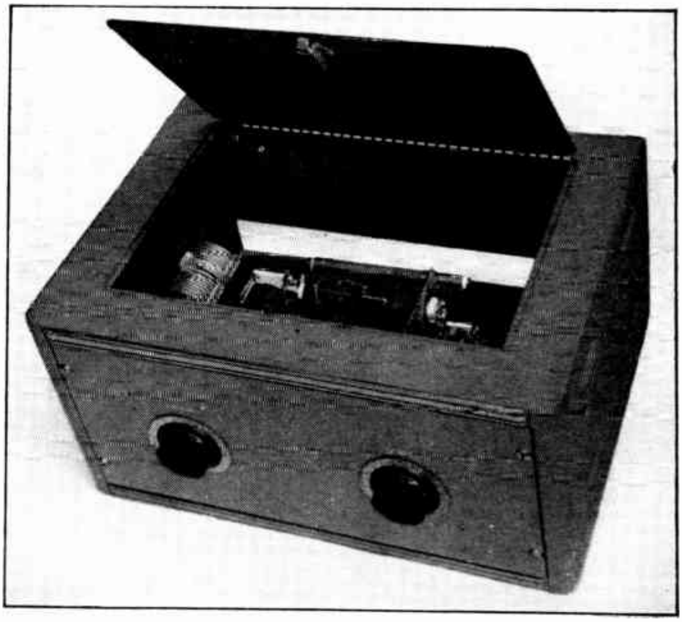


Fig. 6-105 — The push-pull triode amplifier installed in a standard metal cabinet.

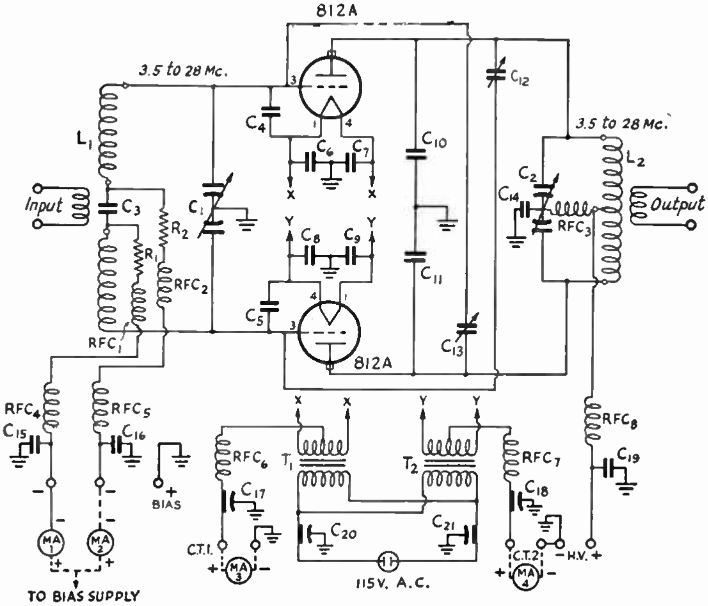


Fig. 6-106 — Circuit diagram of the push-pull triode amplifier.

C_1 — Dual variable — 100 $\mu\mu$fd. per section, 0.03-inch spacing (Hammarlund HFAD-100-B).
C_2 — Dual variable — 100 $\mu\mu$fd. per section, 0.07-inch spacing (Hammarlund HFBD-100-E).
C_3 — 0.001-μfd. mica.
C_4, C_5 — 12-$\mu\mu$fd. ceramic.
C_6, C_7, C_8, C_9 — 0.001-μfd. mica.
C_{10}, C_{11} — Tubular air condenser (see text).
C_{12}, C_{13} — Neutralizing condenser (see text).
C_{14} — 0.001-μfd. 5000-volt-wkg. mica.
C_{15}, C_{16} — 470-$\mu\mu$fd. mica.
C_{17}, C_{18} — 0.1 μfd., 250 volts (Sprague Hypass).
C_{19} — 500-$\mu\mu$fd. 5000-volt-wkg. mica.
C_{20}, C_{21} — 0.01 μfd., 600 volts (Sprague Hypass).
R_1, R_2 — 100 ohms, 1 watt.
L_1, L_2 — See coil table.
MA_1, MA_2 — D.c. milliammeter, 50-ma. scale.
MA_3, MA_4 — D.c. milliammeter, 300-ma. scale.
RFC_1, RFC_2 — 2.5-mh. 125-ma. r.f. choke.
RFC_3 — 1-mh. 600-ma. r.f. choke (National R154U).
RFC_4, RFC_5, RFC_6, RFC_7, RFC_8 — 7-μh. r.f. choke (Ohmite Z-50).
T_1, T_2 — Filament transformer: 6.3 volts, 4 amp. (for 812As).

insulated from the chassis, of course. The jack bar for the plate tank coil is mounted on ½-inch cone insulators, close to the tank condenser.

An aluminum bracket 7 inches long and 3½ inches high holds the two tube sockets, spaced 4½ inches apart. The grid tank condenser also is mounted directly on this bracket without insulation. Metal spacers are used to bring the shaft of the condenser 3½ inches from the end of the chassis to match the shaft of the plate tank condenser. The grid tank-coil socket is placed next to the condenser so that the axis of the coil will be at right angles to that of the plate coil. The National AR-17 grid coils are cut at the center and the open end is connected to a spare pin in the base.

The two fixed air condensers, C_{10} and C_{11}, are made as shown in Fig. 6-104 and are fastened to the tube-socket bracket. A standard 9/16-inch plate grip makes a convenient connection to the tubing and also serves as a means of mounting.

The photograph of Fig. 6-107 shows the construction of the tubular neutralizing condensers. They are quite similar to the tubular fixed condensers, but provision is made for moving the rod in and out of the tubing. A National ceramic button, cemented in the tubing at the control end, serves as a bearing for the rod. The other end of the rod is threaded 6-32 to fit a National GS-1 ceramic pillar which is ½ inch in diameter. This holds the rod central within the tubing as it is slid back and forth. Clearance holes, lined with rubber grommets, are cut in the tube bracket to admit the condenser rods and the rods are provided with sliding

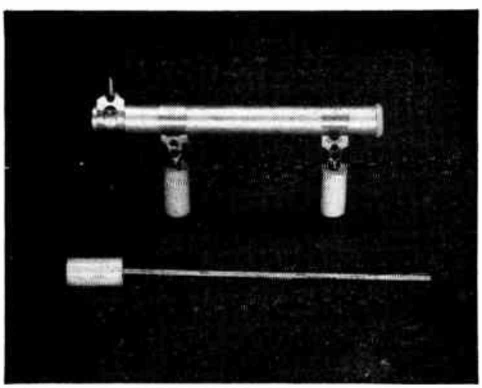


Fig. 6-107 — This photograph shows the construction of the tubular neutralizing condensers for the push-pull triode amplifier. The tubing is ½-inch inside diameter, mounted on ceramic pillars by means of plate grips soldered to screws in the pillars. The inner conductor threads into the end of a third pillar that supports the loose end of the rod.

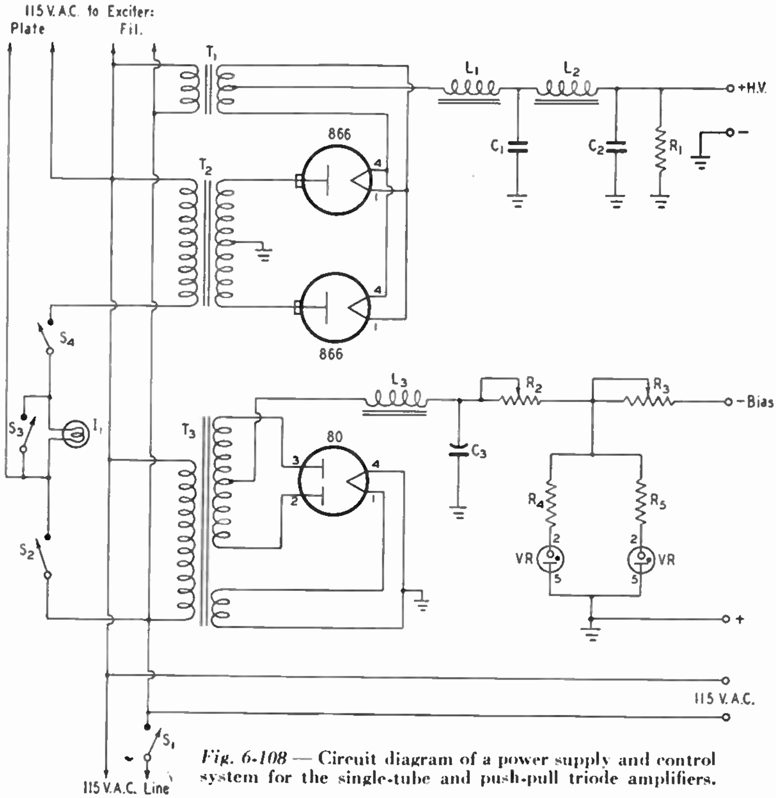


Fig. 6-108 — Circuit diagram of a power supply and control system for the single-tube and push-pull triode amplifiers.

C_1, C_2 — 4-μfd. 2000-volt oil-filled.
C_3 — 8-μfd. 450-volt electrolytic.
R_1 — 25,000 ohms, 150 watts.
R_2 — 50,000 ohms, 25 watts, adjustable.
R_3 — 1500 ohms, 25 watts, adjustable.
R_4, R_5 — 100 ohms, 1 watt.
L_1 — 5/25-hy. swinging choke, 200-ma. for single tube, 400-ma. for push-pull.
L_2 — 20-hy. smoothing choke, 200-ma. for single tube, 400-ma. for push-pull.
L_3 — 30-hy. 50-ma. filter choke.
I_1 — 150-watt 115-volt lamp.
S_1 — 15-amp. switch.
S_2, S_3, S_4 — 10-amp. switch.

S_1 is the main power switch, turning on all filaments and the bias supply and setting up the circuit for S_2 which controls the exciter plate supply. S_2 also sets up the circuit for S_4 which turns on the high-voltage supply. S_3 cuts in or out I_1 for reducing power during transmitter adjustment.

T_1 — Filament transformer: 2.5 volts, 10 amp., 10,000-volt insulation.
T_2 — Plate transformer: 1200/1500 volts d.c., 200 or 400 ma.
T_3 — Power transformer: 650 v. a.c., c.t., 50 ma. or more; 5 volts, 2 amp.
VR — VR-75 voltage regulator (see text).

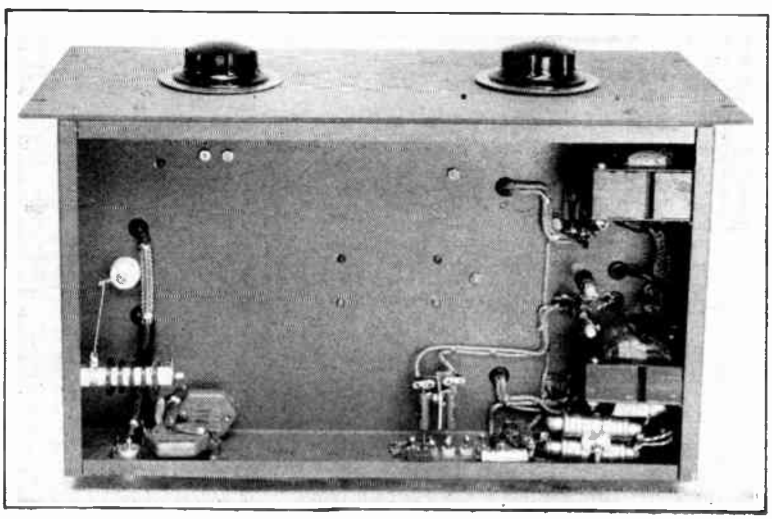


Fig. 6-109 — Bottom view of the push-pull triode amplifier showing the placement of the filament transformers and v.h.f. filter components.

COIL TABLE FOR PUSH-PULL AMPLIFIER

Band	*Turns*	*Wire No.*	*Diam.*	*Length*	*Lμh.*
** Plate*					
27–28 Mc.	4	8	2″	5″	0.6
14 Mc.	10	10	2¼″	2″	3.4
7 Mc.	18	12	3″	2½″	13
3.5 Mc.	26	14	3″	3″	37
*** Grid*					
27–28 Mc.	6	18	1¼″	1″	0.8
14 Mc.	12	18	1¼″	1″	3.6
7 Mc.	28	22	1¼″	1½″	15
3.5 Mc.	56	24	1¼″	1¾″	55

* National AR18 series coils. AR18-6 6-meter coil used for 28 Mc. One turn off each end of coils for other bands.

** National AR17 series coils. One turn off each end on AR17-10 and AR17-20. Other coils unaltered.

contacts (banana jacks) soldered to the tube-socket grid terminals. Insulated control knobs can be fastened to the protruding ends of the rods, but it is safer to adjust the condensers in small steps with excitation off.

The filament transformers and harmonic-filter components are placed underneath the chassis as in the single-tube amplifier. All power wiring is done with shielded wire.

Both units are shown installed in a standard steel cabinet. An enclosure of screening would be equally satisfactory as a shield and might have the advantage of better ventilation. If the cabinet is used, it would be advisable to punch several ventilating holes in the cover immediately above the tubes and others in the chassis below the tubes and in the bottom of the cabinet, since a very considerable amount of heat is generated in the course of operation.

Adjustment

The circuit diagram of a power-supply section and a control system for these amplifiers is shown in Fig. 6-108. With the tubular condensers shown, the plate voltage should be limited to about 1200 volts with plate modulation. An exciter capable of delivering about 25 watts should be provided for the single-tube amplifier and about twice this driving power should be available for the push-pull stage. Fig. 6-106 shows how the dual meters should be connected in the push-pull amplifier. For 812-A tubes, VR-75s may be used in the bias supply. Only a single regulator tube need be used in the case of the single-tube amplifier. In either case, R_2 should be set initially so that at least one of the regulators just ignites. For c.w. operation at 1500 volts, R_3 should be adjusted to 1500 ohms for a single tube or to 750 ohms for two tubes. For plate-modulated 'phone operation at 1200 volts, the grid resistor should be set at 1000 ohms for a single tube, or 500 ohms for two tubes. Grid current should run 30 ma. per tube for c.w. operation and 35 ma. per tube for 'phone operation, both under full load (175 ma. per tube c.w. or 140 ma. per tube 'phone).

For other tube types, and for tuning procedure, see the tube tables, Chapter Twenty-Five, and earlier sections of this chapter.

Unbalance in the cathode currents of the p.p. amplifier can be corrected by adjustment of the length of one or the other of the two plate leads, between the junction of the connection to the tubular condenser and the tank condenser. Greatest unbalance is likely to occur at 28 Mc., so a satisfactory adjustment for this band will usually hold for lower frequencies.

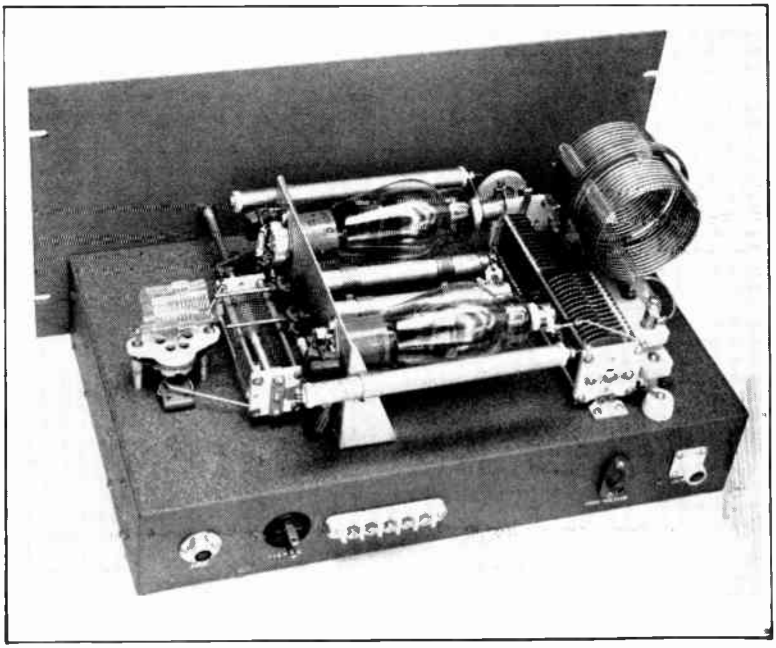


Fig. 6-110 — Rear view of the push-pull triode amplifier. The grid tank condenser is fastened to the bracket holding the tube sockets. The neutralizing condensers are between the tubes with their inner rods extending through holes in the bracket.

A 1-Kw. Beam-Tetrode Amplifier

Figs. 6-111 through 6-115 show the circuit diagram and construction of a single-tube screen-grid amplifier capable of handling up to 1-kw. input on c.w., or 675 watts on plate-modulated 'phone. It is designed to be operated in any band from 80 through 10 meters by the use of plug-in coils. Any exciter capable of delivering 15 to 20 watts should provide adequate excitation for the 4-250A in this amplifier.

The circuit diagram is shown in Fig. 6-113. It is a conventional link-coupled arrangement except for the inductive link neutralizing system (L_2 and L_4). This neutralization is desirable to maintain reliable stability on all bands. All power leads are filtered for v.h.f. harmonics.

Construction

The amplifier is designed for use in a standard rack cabinet or other shielding enclosure. To that end, it is arranged so that both grid and plate coils may be removed by pulling toward the rear. Thus the chassis is inverted to provide access to the grid coil.

On top, the plate tank condenser is inverted and mounted with metal angles on 2-inch ceramic cone insulators. It is placed so that its shaft will come at the center. The jack bar for the tank coil is fastened between an angle piece at the forward end of the tank-condenser frame and another angle piece bolted to one of the panel brackets. The mounting is made so that the coil is tilted at an angle of about 45 degrees. The antenna-coupling link shaft is driven from a control on the panel by means of a Millen right-angle gear box. The neutralizing link, L_4, is the B & W type BVL. The assembly is fastened with a single screw to the top of a 1¼-inch ceramic pillar mounted on the rear corner of the tank condenser. This mounting permits the link to be pivoted on the pillar as well as hinged in the usual fashion.

Since coils with a variable end link are not available, center-link coils have been adapted to the purpose by using only one section of the two-section coils. As a matter of convenience in changing bands, the unused section of one coil is removed and a section of coil for an adjacent band is mounted instead. Thus each coil plug strip carries coils for two bands and the change from one to the other is made simply by turning the unit end for end. The two unused jacks in the jack bar are connected together with a copper strap so that the unused section of coil is short-circuited.

The tube socket (National HX-100-S) is submounted alongside the rear corner of the tank condenser where the plate lead to the stator terminal can be made short. The filament transformer is mounted at the front of the chassis out of the direct field of the tank. A clearance hole is cut for the terminals which protrude underneath.

Underneath, the grid tank condenser is mounted at the center of the chassis on ½-inch stand-off insulators. A 3⅜-inch strip of aluminum is bent as shown in the bottom-view photograph to form a mounting for the grid tank coil directly to the rear of the condenser, as well as a shielding enclosure for the components in the power-lead filters. The leads between the coil socket and the condenser pass through small bushings (Millen 32150) or clearance holes in the aluminum. The socket and ventilating fan are enclosed in a 6 × 4 × 3¼-inch box made of aluminum sheet. When the bottom plate of the box is in place, the fan forces air up through the socket to the tube. The box should be perforated with ¼-inch holes back of the fan to provide an air intake.

The filament, screen and grid by-pass condensers are mounted directly at the tube socket. All are grounded at the same point —

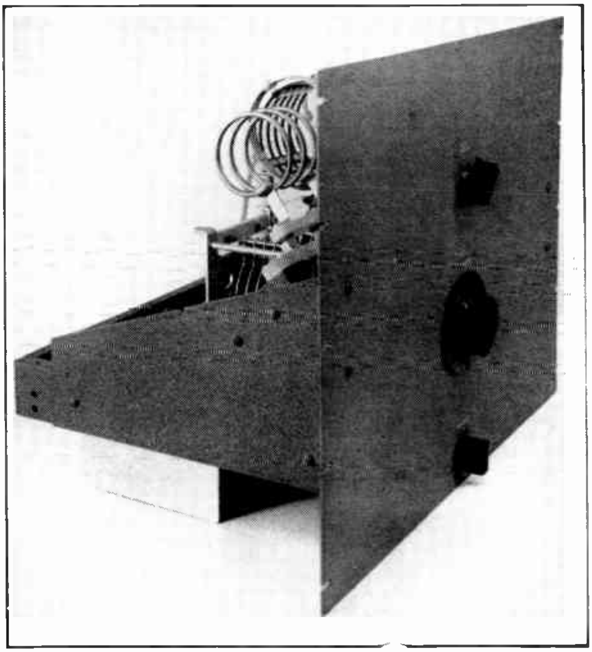


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Fig. 6-111 — Front view of the 4-250A amplifier, showing the method of assembling the panel and the chassis. The controls on the panel, from top to bottom, are the output coupling knob, plate tuning dial, and grid tuning dial. The panel is 19 by 17½ inches.

one of the socket mounting screws. A ceramic terminal strip for the a.c. line, bias, screen voltage and ground terminals, a Millen safety terminal for the plate-voltage connection, and a coaxial jack for r.f. input are mounted on the rear surface of the shielding strip.

All power wiring is done with shielded wire. The high-voltage lead is a length of high-voltage ignition cable covered with ½-inch shielding braid up to within an inch of each end.

The grid-circuit neutralizing link consists of two turns of No. 14 wire, 1½ inches in diameter, fastened to a pair of 2½-inch pillar insulators (National GS-2) so that the coil is coupled to the low-potential end of L_1 and yet does not interfere with the removal of the grid coil.

Plate-Coil Modification

The 80-meter HDVL coil is dismounted from its ceramic plug bar and a diagonal cut is sawed through the center of the plastic strip holding the two sections of the coil. The 40-meter coil is similarly cut. One section of the 80-meter coil and one section of the 40-meter coil are then reassembled as a unit by cementing together at the center, the diagonal cuts overlapping. The coils for 14- and 28-Mc. operation are altered in the same way. Other combinations may be made up as desired, depending upon the bands wanted. The 21-Mc. coil may be a separate unit or combined with the coil for another band.

Adjustment

The circuit diagram of a suitable power-supply unit for this amplifier is shown in Fig. 6-114. Caution should be exercised in operating a beam tetrode with fixed screen supply — especially a high-power tube — since the screen current in the absence of plate voltage and full load can run to damaging limits.

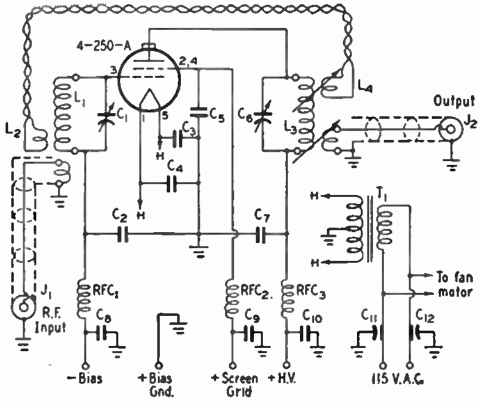

Fig. 6-113 — Schematic diagram of the 4-250A amplifier.
C_1 — 100-μμfd. variable (National TMS-100).
C_2, C_3, C_4 — 0.0022-μfd. mica.
C_5 — 0.001-μfd. 1000-volt mica.
C_6 — 150-μμfd. 6000-volt peak (National TMA-150-A).
C_7 — 0.001-μfd. 5000-volt-wkg. mica.
C_8 — 470-μμfd. mica.
C_9 — 500-μμfd. 1000-volt mica.
C_{10} — 500-μμfd. 5000-volt-wkg. mica.
C_{11}, C_{12} — 0.005 μfd., 600 volts (Sprague Hypass).
L_1 — Millen 43000 series coils:
 3.5 Mc. — 32 turns No. 20, 1½-in. diam., 1½ in. long, 7-turn link (43082 with 6 turns removed).
 7 Mc. — 24 turns No. 16, 1½-in. diam., 2 in. long, 7-turn link (43042).
 14 Mc. — 9 turns No. 16, 1½-in. diam., 1½ in. long, 2-turn link (43022).
 21–28 Mc. — 4 turns No. 14, 1½-in. diam., 1⅜ in. long, 2-turn link (43012).
L_2 — 2-turn link, No. 14, 1½ inches diam.
L_3 — B & W HDVL series (modified, see text).
 3.5 Mc. — 16 turns No. 10, 3½-in. diam., 3 in. long.
 7 Mc. — 10 turns No. 8, 3½-in. diam., 2⅞ in. long.
 14 Mc. — 6 turns No. 8, 3½-in. diam., 3 in. long.
 21 Mc. — 4 turns 3/16-in. copper tubing, 3-in. diam., 2⅞ in. long.
 28 Mc. — 3 turns 3/16-in. copper tubing, 2⅜-in. diam., 2⅝ in. long.
L_4 — 3-turn swinging link, No. 18, 2⅝-in. diam., ¼ in. long (BVL link assembly).
J_1, J_2 — Coaxial connector (Amphenol 83-1R).
RFC_1, RFC_2, RFC_3 — 7-μh. r.f. choke (Ohmite Z-50).
T_1 — Filament trans.: 5 volts, 14.5 amp. (UTC S-59).

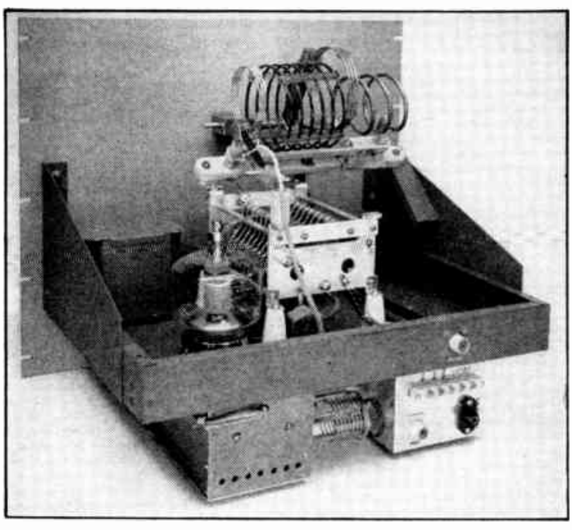

Fig. 6-112 — Rear view of the 4-250A amplifier. The construction of the reversible plug-in plate coil is shown. The small variable link at the left of the plate coil is a part of the neutralizing circuit. The grid-coil compartment is seen below the chassis between the shield box that houses the fan and the partition on which the input terminals are mounted.

It is advisable to conduct all preliminary adjustments at reduced screen voltage to keep the screen dissipation at a safe level. The lamps, I_1 and I_2 in Fig. 6-114, are for this purpose. A size of lamp should be selected that will give the desired reduction in screen and plate voltage, remembering that the lamps with lower wattage rating have a higher resistance and therefore will give a greater voltage reduction.

Neutralization is merely a matter of adjusting the position of the plate neutralizing link for complete stability. Since the system depends upon correct polarization of the links, it may be necessary to reverse the connections to *one* of the links.

For operating at a plate voltage of 3000, normal excitation is indicated when the grid current is 10 ma. and the bias 180 volts with the amplifier loaded to draw a plate current of 325 ma. Under these conditions, the screen current with a screen-supply voltage of 500 should run approximately 60 ma. For plate-modulated 'phone operation at 3000 volts, the grid current should be 9 ma. at 310 volts under full load and the screen current 30 ma. at 400 volts. Under the above conditions, R_3, Fig. 6-114, should be set at 3000 ohms for c.w. operation and at 18,000 ohms for 'phone operation. R_2 should be adjusted so that the VR tube just ignites without excitation.

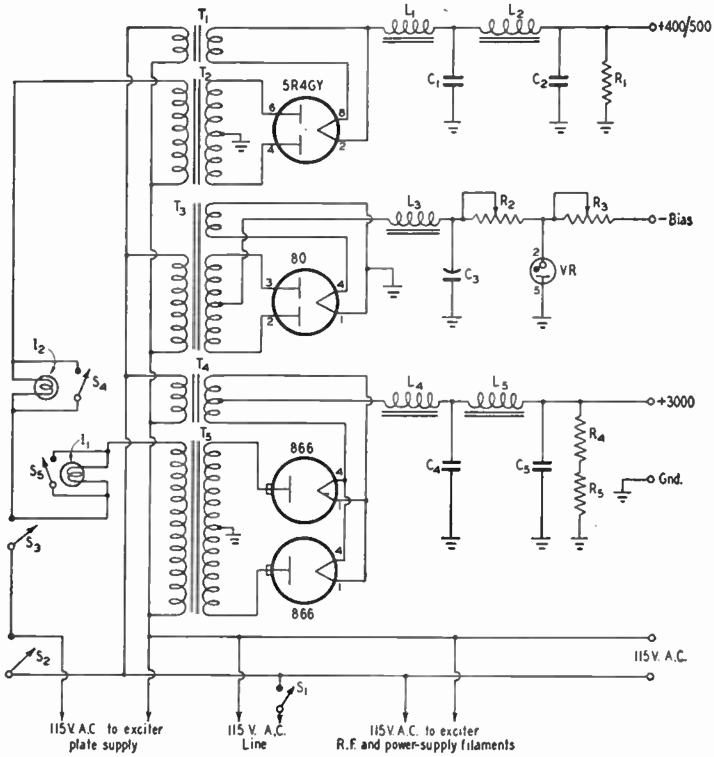

Fig. 6-114 — Circuit diagram of a power supply for the beam-tetrode amplifier. S_1 is the main switch, turning on all filaments. S_2 turns on the plate voltage for the exciter unit and sets up the circuit for S_3 which turns on both screen and plate supplies for the amplifier. I_1 and I_2 are 115-volt lamps of proper size to reduce screen and plate voltages to a suitable value for tuning. S_4 and S_5 short-circuit these lamps for full-power operation.

C_1, C_2 — 1-µfd. 600-volt oil-filled.
C_3 — 1-µfd. 450-volt-wkg. electrolytic.
C_4, C_5 — 4-µfd. 3000-volt oil-filled.
R_1 — 25,000 ohms, 25 watts.
R_2 — 50,000 ohms, 25 watts, adjust.
R_3 — 20,000 ohms, 25 watts, adjustable.
R_4, R_5 — 50,000 ohms, 75 watts.
L_1, L_2 — 20-hy. 100-ma. filter choke.
L_3 — 30-hy. 50-ma. filter choke.
L_4 — 5/25-hy. 400-ma. swinging choke.
L_5 — 20-hy. 400-ma. smoothing choke.
I_1, I_2 — Power-reducing lamp.
S_1 — 15-amp. switch.
S_2, S_3, S_4, S_5 — 10-amp. switch.
T_1 — Filament transformer: 5 volts, 2 amp.
T_2 — Plate transformer: 500 volts d.c., 100 ma.
T_3 — Power transformer: 250–350 volts d.c., 75 ma.; 5 volts, 3 amp.
T_4 — Filament transformer: 2.5 volts, 10 amp., 10,000 volts insulation.
T_5 — Plate transformer: 3000 volts d.c., 400 ma.
VR — Voltage regulator — VR-150.

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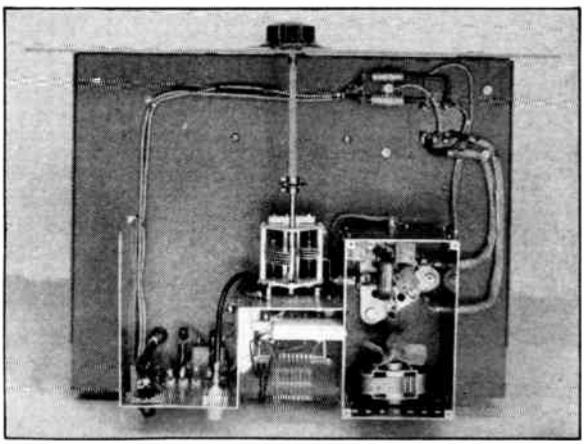


Fig. 6-115 — Bottom view of the 4-250A amplifier. The harmonic filters are in the compartment to the left of the grid coil. The arrangement of the by-pass condensers inside the fan housing is also shown, with the grid terminal of the tube socket pointing toward the grid tank circuit. The chassis measures 17 by 13 by 2 inches.

A 20-Watt Mobile Transmitter for 40 and 80 Meters

The circuit diagram of a 20-watt transmitter for 75-meter 'phone and 40- or 80-meter c.w. is shown in Fig. 6-118. A modified Pierce crystal oscillator drives a 2E26 amplifier. Both bands are covered by the oscillator tank condenser, C_5, with a single coil at L_1. The output circuit of the 2E26 is in the form of a pi-section network which permits feeding power into antennas of a variety of lengths. A ceramic tap switch, S_1, adjusts the antenna inductance for each band.

Fig. 6-120 shows the circuit of the modulator. The two sections of the 12AU7 dual triode serve as speech amplifier and driver for the 6X7 Class B modulator. The audio section is designed for use with a low-output single-button carbon microphone such as the T-17. A 3-section 3-pole rotary switch, S_1, Fig. 6-120, performs the necessary change-over functions in shifting the rig between 'phone and c.w. operation.

Construction

The transmitter unit is made up of two separate sections housed in a standard 9 × 5 × 6-inch steel utility box. The r.f. unit is assembled on an aluminum chassis 6 inches long, 1¼ inches wide and 1 inch deep. All parts in the r.f. unit, with the exception of the output coil and tap switch, are mounted on the chassis. The switch is fastened to the box cover, which serves as the panel, and the coil is mounted directly behind, supported at one end by a 2-inch ceramic pillar and at the other end by a short length of heavy wire that extends from the feed-through antenna terminal to the tuning condenser below it.

Parts underneath the chassis should be placed carefully so that they will not interfere with the audio components below. The oscillator plate coil and tuning condenser are mounted along one edge where they extend down into the space just above the driver transformer and the 12AU7 on the audio chassis. The smaller parts in the r.f. unit are mounted near the other edge, as close to the chassis as possible to insure adequate clearance for the modulation transformer and the 6X7 on the left-hand side.

The audio unit is built on a similar chassis having the same dimensions but with ¼-inch

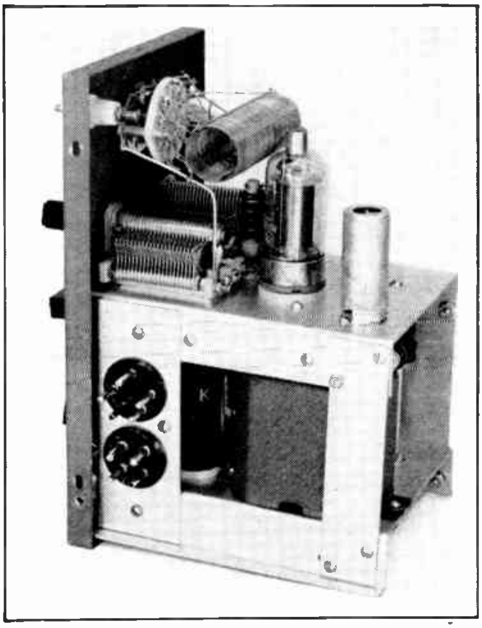


Fig. 6-117 — General view of the complete transmitter assembly. The r.f. circuits occupy the upper deck, and are held above the audio section by simple tab brackets. Power and control cables enter through connectors mounted on one of these tabs. The construction of the plate circuit of the 2E26 stage is shown, with the r.f. choke just visible between the two tuning condensers.

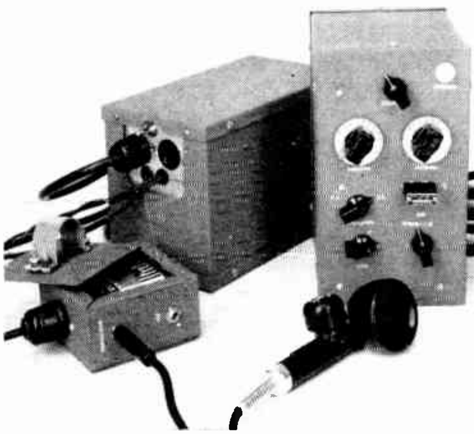


Fig. 6-116 — A compact 20-watt mobile transmitter for the 3.5- and 7-Mc. bands. Designed for mounting under the dashboard of the car, the transmitter and its power supply occupy standard 5 × 6 × 9-inch utility boxes, and the control box, which clamps to the steering post, is in a 4 × 4 × 2-inch box.

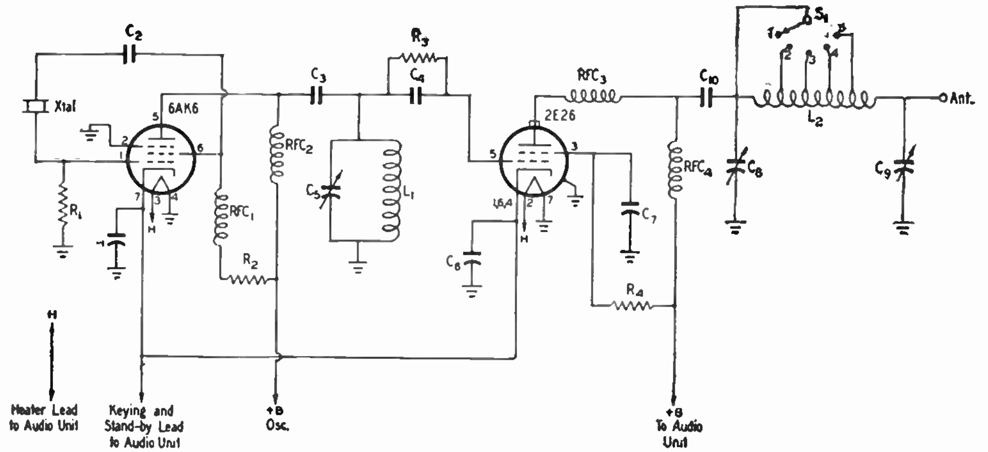


Fig. 6-118 — Circuit diagram of the 20-watt mobile r.f. unit

C_1, C_6 — 0.01-μfd. paper, 600 volts.
C_2, C_{10} — 0.001-μfd. mica, 500 volts.
C_3 — 0.0047-μfd. mica, 500 volts.
C_4 — 47-μμfd. mica, 500 volts.
C_5 — 250-μμfd. variable (National STH-250).
C_7 — 0.0068-μfd. mica, 500 volts.
C_8, C_9 — 335-μμfd. variable (National STH-335).
R_1 — 47,000 ohms, ½ watt.
R_2 — 56,000 ohms, 1 watt.
R_3 — 22,000 ohms, 1 watt.
R_4 — 15,000 ohms, 10 watts.
L_1 — 32 turns No. 22 enam. close-wound on ¾-inch diam. form.
L_2 — 48 turns No. 20 tinned, 3⅛ inches long, 1-inch diam., taps at 12, 22, 32 and 40 turns from plate end (B & W Miniductor No. 3015).
RFC_1, RFC_2, RFC_4 — 2.5-mh. 100-ma. r.f. choke (National R-100-S).
RFC_3 — 25 turns No. 26 enam. close-wound on ¼-inch diam. form (National R-33, 1 μh.).
S_1 — Single-pole 5-position ceramic switch.

lips bent out along the bottom to provide rails on which the assembly rides when it is being slipped into the box. The parts beneath the audio chassis are mounted as close to the chassis as possible.

Aluminum strips 4½ inches long by 1 inch wide serve as braces at the rear between the two chassis. A similar strip 1⅝ inches wide, at the front on the right side, provides a mounting for the two connectors used to bring the supply voltage and the control circuits into the unit. A cut-out is made in the edge of the box to clear these connectors.

TYPICAL OPERATING DATA — 20-WATT MOBILE TRANSMITTER

Conditions: c.w.; loaded to 110 ma. total cathode current; supply voltage (under load) 390 volts; 80-meter crystal used.

Stage	*80-Meter Output*		*40-Meter Output*	
	Volts	*Ma.*	*Volts*	*Ma.*
6AK6				
plate	390	19	390	21
screen	200	3	210	3.5
2E26				
plate	390	78	390	78
screen	200	6	210	5
grid *	— 100	4	— 90	3

* Grid current and voltage will vary widely from these figures depending on tuning. Optimum obtainable values are shown.

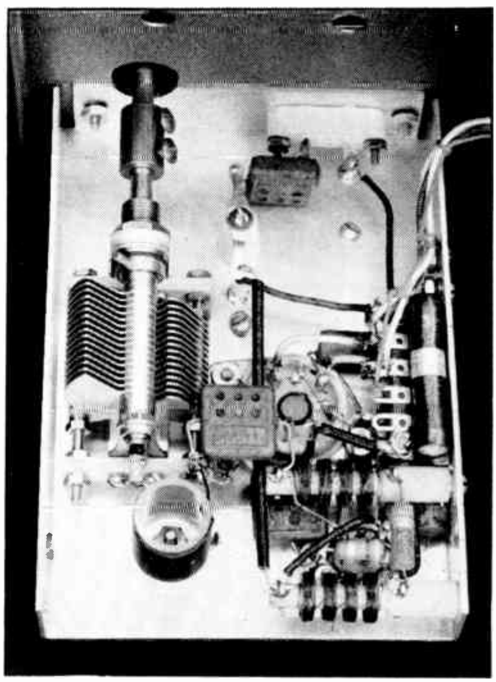


Fig. 6-119 — Bottom view of the r.f. chassis. The oscillator socket is partly hidden from view by the two r.f. chokes that are mounted on the right-hand chassis edge. Directly to the left of the oscillator socket is the oscillator plate coil. The socket for the 2E26 is mounted a little to the right of center of the chassis, close to both the oscillator socket and the tuning condenser.

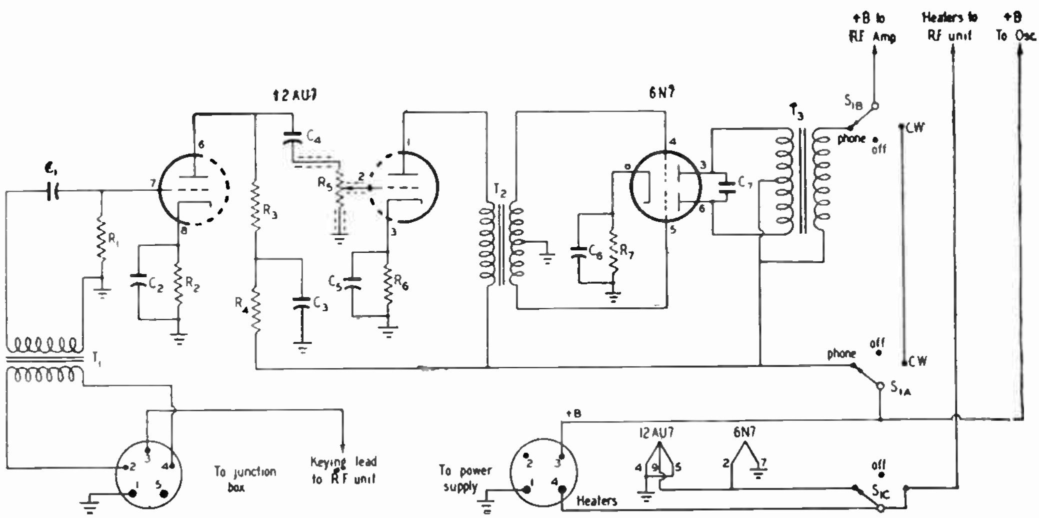


Fig. 6-120 — Circuit of modulator for 20-watt mobile transmitter.

C_1 — 0.1-μfd. paper.
C_2, C_5 — 10-μfd. 25-volt electrolytic.
C_3 — 8-μfd. 450-volt electrolytic.
C_4 — 0.01-μfd. paper.
C_6 — 50-μfd. 50-volt electrolytic.
C_7 — 0.0068-μfd. mica, 500 volts.
R_1 — 0.47 megohm, ½ watt.
R_2 — 2200 ohms, ½ watt.
R_3 — 0.1 megohm, ½ watt.
R_4 — 4700 ohms, 1 watt.
R_5 — 0.5 megohm potentiometer, audio taper.
R_6 — 560 ohms, ½ watt.
R_7 — 220 ohms, 2 watts.
S_1 — 3-pole 3-position rotary switch.
T_1 — Midget microphone transformer, s.b. mic. to grid (Inca F-65).
T_2 — Driver transformer, single plate to Class B grids (Thordarson T-20D76).
T_3 — Multitap modulation transformer (UTC S-18, connected to match 8000-ohm primary to 4000-ohm secondary).

Power Supply

The power-supply circuit is shown in Fig. 6-122. A combination transformer is used to permit operation from either the 115-volt a.c. line or a 6-volt car battery. Two 6X5 rectifiers in parallel are used to carry the total current drain of about 120 ma. Hash filtering is accomplished by chokes and by-pass condensers. Separate output connectors are used for 6- and 115-volt power sources. When operating from a battery, d.c. is applied directly to the heaters. When a.c. input is used, the other connector supplies 6.3 volts a.c. to all heaters from the transformer. A s.p.d.t. switch, S_2, shifts the heaters of the 6X5s. All parts in the power supply are mounted on a 4¼ × 8¾ × 1-inch aluminum chassis. The entire supply is enclosed in a second utility box of the same size as that used for the transmitter.

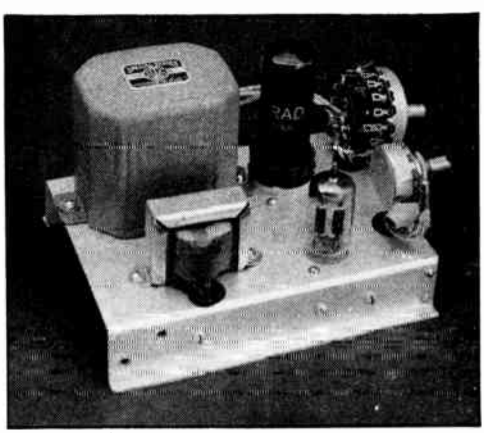


Fig. 6-121 — Top view of the audio unit. At the right-hand edge are the gain control and the 'phone-c.w. switch. The 12AU7 and the 6N7 are mounted in line behind the two controls. The transformers occupy the rear of the chassis, located in such position that they clear all parts in the r.f. unit, which mounts above them.

Control Circuits

The circuit of the control box is shown in Fig. 6-123. The box is a standard item 4 by 4 by 2 inches, with a bracket bolted to one of the covers for clamping to the steering wheel. Jacks for both microphone and key and a 4½-volt microphone battery are included. A 5-terminal receptacle is mounted on the bottom of the box to bring the control cable into the box from the transmitter unit.

Only two cables are required. One is a 3-wire shielded cable that runs from the control box to the transmitter. The other requires 3 conductors, one for high voltage, one for heater voltage and the third for ground. The ground and filament leads should be made of as heavy wire as possible to minimize voltage drop.

The control circuit is arranged so that push-to-talk or break-in operation is possible. The switch on the microphone controls the transmitter, once the main power switch has been turned on. In 'phone operation, the microphone switch closes the cathode circuits of the two tubes in the r.f. section. In the "stand-by" position the cathode circuits are opened. Plate voltage is still applied to the audio circuits but the microphone circuit is opened. In c.w. operation, the 'phone-c.w. switch removes plate voltage from the audio tubes.

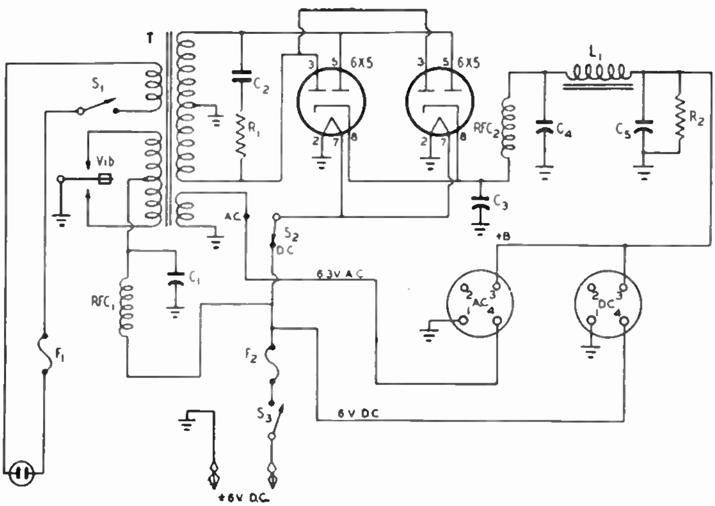


Fig. 6-122 — Circuit diagram of the power supply used with the 20-watt mobile rig. Provisions are made for operation from either the 115-volt a.c. line or from a 6-volt storage battery.

C_1 — 0.5-µfd. paper, 50 volts or more.
C_2 — 0.005 µfd., 1600 volts.
C_3 — 0.01-µfd., 600 volts.
C_4 — 8-µfd., 450 volts, electrolytic.
C_5 — 32-µfd., 450 volts, electrolytic (dual 16-µfd. condenser with sections in parallel).
R_1 — 4700 ohms, 1-watt carbon.
R_2 — 25,000 ohms, 20 watts, wire-wound.
L_1 — 2.5 hy., 100-ma. filter choke, 100 ohms d.c. resistance (Stancor C-2303).
F_1 — 2-amp. fuse.
F_2 — 15-amp. fuse.
RFC_1 — 11 turns No. 14 enameled, ½-inch diam., 2½ inches long.
RFC_2 — 2.5 mh., 300 ma. (National R-300).
S_1 — S.p.s.t toggle switch.
S_2 — S.p.d.t. toggle switch.
S_3 — Heavy-duty s.p.s.t. toggle switch.
T — 6-volt vibrator transformer, with separate 115-volt primary, 350-0-350 v. r.m.s., 125 ma., and 6.3 v. at 2.25 amp. (Stancor P-6166).

By using standard steel angle brackets, the transmitter may be supported underneath the dashboard, against the bulkhead, or any place where it will fit. The control box mounts on the steering wheel and the power supply should be placed as close to the battery as possible. The pi-section output circuit will permit full loading even when a short length of wire or a whip antenna is used.

Adjustment

The accompanying table shows typical voltage and current readings for the two bands.

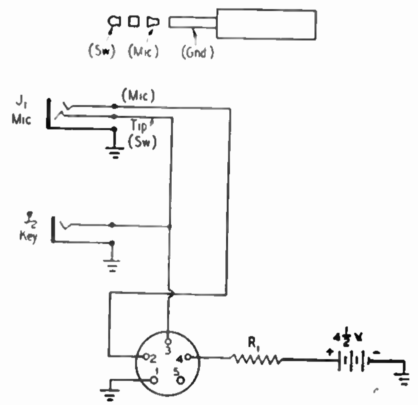


Fig. 6-123 — Schematic diagram of the control box. Connections of the 2-circuit plug used with the T-17-B microphone are shown in the sketch at the top.

J_1 — 2-circuit microphone jack.
J_2 — Open-circuit key jack.
R_1 — 220 ohms, 1 watt.

No special provision has been made for meters in the unit, since the transmitter can be tuned up with a milliammeter plugged into the key jack.

The oscillator tank circuit should tune to the 3.5-Mc. band near maximum capacitance of the tank condenser and to 7 Mc. near minimum. Resonance will be indicated by an upward kick in the reading of the milliammeter. With one of the condensers in the pi-section filter set at about half capacitance, the other should be adjusted for a dip in plate current with the oscillator running. If this dip cannot be found, the switch should be turned to a different tap and the process repeated. Connection of the antenna will cause a change in the tuning of the output circuit. Both taps and condensers should be readjusted until resonance as indicated by the dip in cathode current is found again. If the current at resonance is too low, the capacitance in the two condensers should be reapportioned, increasing one and decreasing the other to restore resonance, or vice versa, until the total current at resonance is about 100 ma.

When tuning for 3.5-Mc. output, the output should be checked with an absorption wavemeter to make sure that the 2E26 final is not doubling frequency.

Rack Construction

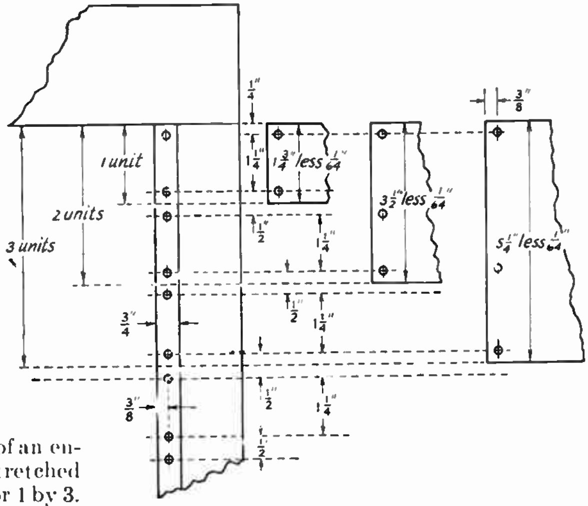


Fig. 6-125 — Detail sketch showing proper drilling for standard rack and panels. As shown for the 3½- and 5¼-inch panels, only sufficient holes are drilled in the panel to provide the necessary strength. When the panels are drilled as shown, they may be moved up and down in steps of 1¾ inches and the holes will always match.

Many of the units described in the constructional chapters of the *Handbook* are designed for a standard rack mounting. This standardization facilitates the assembly and modification of station equipment. Since the advent of television, racks of the enclosed type have become a matter of practical necessity for transmitters to be operated without interference in neighborhoods where television receivers are in use. While enclosed cabinet-type racks of metal are available on the market, many amateurs prefer to build their own less expensively from wood and copper screening. With care, an excellent substitute can be made.

Fig. 6-124A shows a broken top view of an enclosed rack made of copper screening stretched over a framework of wood strips 1 by 2 or 1 by 3. The copper screen, represented by the dashed lines and the cross-hatching, is stretched over the outside of each frame, wrapped around the ends on all four sides and tacked fast on the inside. The top and bottom are made in similar fashion. When the frames are fastened together, the screening makes contact all along each joint. Contact at the hinge of the door at the rear is assured by the use of a full-length piano hinge. Trim strips of thin wood along the two vertical 1 by 3s, which hold the panels, and across the top and bottom headers cover up the ragged edges of screening.

As shown in Fig. 6-124B, the panel clearance should be 19$\frac{1}{16}$ inches and the hole centers 18¼ inches apart. Standard panels are in unit heights of 1¾ inches and the hole spacing alternates between ½ inch and 1¼ inches as shown in Fig. 6-125. The table shows the standard drilling for panels of various sizes.

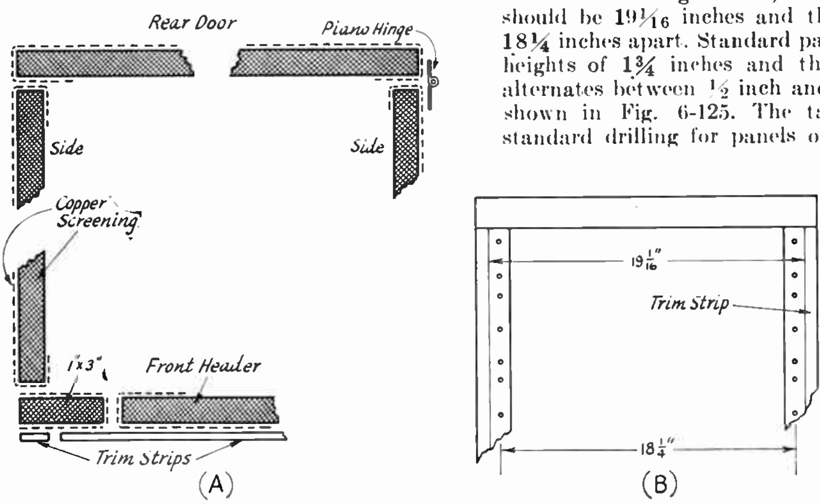


Fig. 6-124 — A — Top detail view of an enclosed relay rack made of wood strips and copper screening. B — Panel-mounting dimensions.

TABLE OF STANDARD RACK DRILLING

Panel Ht. In.	**Holes In.*	*Panel Ht. In.*	**Holes In.*	*Panel Ht. In.*	**Holes In.*	*Panel Ht. In.*	**Holes In.*	*Panel Ht. in.*	**Holes In.*	*Panel Ht. In.*	**Holes In.*
31½	31¼–30	26¼	26 –24¾	21	20¾–19½	15¾	15½–14¼	10½	10¼–9	5¼	5 –3¾
29¾	29½–28¼	24½	24¼–23	19¼	19 –17¾	14	13¾–12½	8¾	8½–7¼	3½	3¼–2
28	27¾–26½	22¾	22½–21¼	17½	17¼–16	12¼	12 –10¾	7	6¾–5½	1¾	1½–¼

* Any or all holes for smaller panels that follow may be added or substituted as desirable. Hole distances are from either top or bottom edges of panel.

CHAPTER 7

Power Supplies

Essentially pure direct-current plate supply is required for receivers to prevent hum in the output. Government regulations require the use of d.c. plate supply for transmitters to prevent modulation of the carrier by the supply, which would result in undesired hum in the case of voice transmissions and an unnecessarily broad c.w. signal.

The filaments of tubes in a transmitter may be operated from a.c. Those in a receiver, excepting the power audio tubes, may be a.c. operated only if the cathodes are indirectly heated.

The comparatively high cost and inconvenience of batteries and d.c. generators preclude their use except where commercial a.c. lines are not available. Wherever such lines are available, it is universal practice to obtain low a.c. voltage for filaments and heaters from a step-down transformer, and the required high-voltage d.c. by means of a transformer-rectifier-filter system. Such a system is shown in the block diagram of Fig. 7-1. Power from the a.c. line is fed to a transformer which steps the voltage up to that required. The stepped-up voltage is changed to pulsating d.c. by passing through a rectifier — usually of the vacuum-tube type. The pulsations then are smoothed out to the required extent by a filtering system.

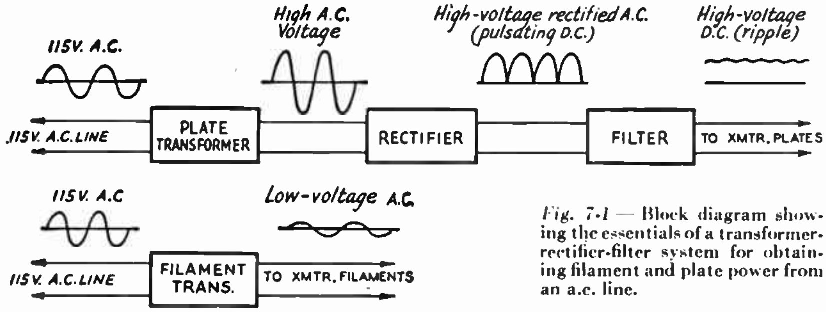


Fig. 7-1 — Block diagram showing the essentials of a transformer-rectifier-filter system for obtaining filament and plate power from an a.c. line.

Rectifier Circuits

Half-Wave Rectifier

Fig. 7-2 shows three rectifier circuits covering most of the common applications in amateur equipment. Fig. 7-2A is the circuit of a half-wave rectifier. During that half of the a.c. cycle when the rectifier plate is positive with respect to the cathode, current will flow through the rectifier and load. But during the other half of the cycle, when the plate is negative with respect to the cathode, no current can flow. The shape of the output wave is shown at the right. It shows that the current always flows in the same direction but that the flow of current is not continuous and is pulsating in amplitude.

The average output voltage — the voltage read by the usual d.c. voltmeter — with this circuit is 0.45 times the r.m.s. value of the a.c. voltage delivered by the transformer secondary. Because the frequency of the pulses in the output wave is relatively low, considerable filtering is required to provide adequately smooth d.c. output, and for this reason this circuit is usually limited to applications where the current involved is small, such as in supplies for cathode-ray tubes and for protective bias in a transmitter.

Full-Wave Center-Tap Rectifier

The most universally-used rectifier circuit is shown in Fig. 7-2B. Being essentially an arrangement in which the outputs of two half-wave rectifiers are combined, it makes use of both halves of the a.c. cycle. A transformer with a center-tapped secondary, or two identical transformers with their secondaries connected in series (with proper polarization), is required with the circuit. When the plate of rectifier No. 1 is positive, current flows through the load to the center-tap. Current cannot flow through rectifier No. 2 because at this instant its cathode is positive in respect to its plate. When the polarity reverses, rectifier No. 2 conducts and current again flows through the

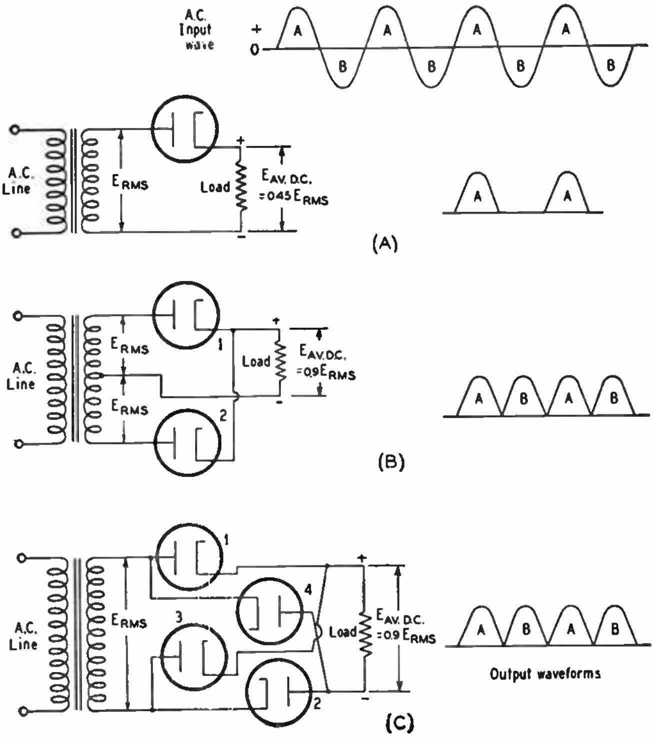


Fig. 7-2 — Fundamental vacuum-tube rectifier circuits. A — Half-wave. B — Full-wave. C — Bridge.

load to the center-tap, this time through rectifier No. 2.

The average output voltage is 0.9 times the r.m.s. value of the voltage across *half* of the transformer secondary. For the same *total* secondary voltage, the average output voltage will be the same as that delivered with a half-wave rectifier. However, as can be seen from the sketch of the output waveform, the frequency of the output pulses is twice that of the half-wave rectifier. Therefore much less filtering is required. Since the rectifiers work alternately, each handles half of the average load current. Therefore the load current which may be drawn from this circuit is twice the rated load current of a single rectifier.

Full-Wave Bridge Rectifier

Another full-wave rectifier circuit is shown in Fig. 7-2C. In this arrangement, two rectifiers operate in series on each half of the cycle, one rectifier being in the lead to the load, the other being in the return lead. Over that portion of the cycle when the upper end of the transformer secondary is positive with respect to the other end, current flows through rectifier No. 1, through the load and thence through rectifier No. 2. During this period current cannot flow through rectifier No. 4 because its plate is negative with respect to its cathode. Over the other half of the cycle, current flows through rectifier No. 3, through the load and thence through rectifier No. 4. The crossover connection keeps the current flowing in the same direction through the load. The output waveshape is the same as that from the simple center-tap rectifier circuit. The output voltage obtainable with this circuit is 0.9 times the r.m.s. voltage delivered by the transformer secondary. For the same total transformer-secondary voltage, the average output voltage when using the bridge rectifier will be twice that obtainable with the center-tap rectifier circuit. However, when comparing rectifier circuits for use *with the same transformer*, it should be remembered that the *power* which a given transformer will handle remains the same regardless of the rectifier circuit used. If the output voltage is doubled by substituting the bridge circuit for the center-tap rectifier circuit, only half the rated load current can be taken from the transformer without exceeding its normal rating. The value of load current which may be drawn from the bridge rectifier circuit is twice the rated d.c. load current of a single rectifier.

Rectifiers

Cold-Cathode Rectifiers

Tube rectifiers fall into three general classifications as to type. The cold-cathode type of rectifier is a diode which requires no cathode heating. Certain types will handle up to 350 ma. at 200 volts d.c. output. The internal voltage drop in most types lies between 60 and 90 volts. Rectifiers of this kind are produced in both half-wave (single-diode) and full-wave (double-diode) types.

High-Vacuum Rectifiers

High-vacuum rectifiers depend entirely upon the thermionic emission from a heated cathode and are characterized by a relatively high internal resistance. For this reason, their application usually is limited to low power, although there are a few types designed for medium and high power in cases where the relatively high internal voltage drop may be tolerated. This high internal resistance makes them less susceptible to damage from temporary overload and they are free from the bothersome electrical noise sometimes associated with other types of rectifiers.

Some rectifiers of the high-vacuum full-wave type in the so-called receiver-tube class will handle up to 250 ma. at 400 to 500 volts d.c. output. Those in the higher-power class can be used to handle up to 500 ma. at 2000 volts d.c. in full-wave circuits. Most low-power high-vacuum rectifiers are produced in the full-wave type, while those for greater power are invariably of the half-wave type.

Mercury-Vapor Rectifiers

In mercury-vapor rectifiers the internal resistance is reduced by the introduction of a small amount of mercury which vaporizes under the heat of the filament, the vapor ionizing upon the application of voltage. The voltage drop through a rectifier of this type is practically constant at approximately 15 volts regardless of the load current. Tubes of this type are produced in sizes that will handle any voltage or current likely to be encountered in amateur transmitters. For high power they have the advantage of cheapness. Rectifiers of this type, however, have a tendency toward a type of oscillation which produces noise in near-by receivers. This can usually be eliminated by suitable filtering.

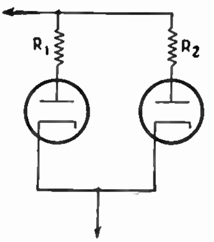


Fig. 7-3 — Connecting rectifiers in parallel for heavier currents. R_1 and R_2 should have the same value, between 50 and 100 ohms.

Selenium Rectifiers

Selenium rectifiers are now available which make it possible to design a power supply capable of delivering up to 400 or 450 volts, 200 ma. These units have the advantage of compactness as well as low internal voltage drop (about 5 volts). However, to limit the charging current with condenser input, a resistance of 25 to 100 ohms should be used in series with the rectifier. They may be substituted in any of the basic circuits shown in Fig. 7-2, the terminal marked "+" or "cathode" corresponding to the cathode in these circuits. Circuits in which the selenium rectifier is particularly adaptable are shown later in Figs. 7-23 through 7-25. Since they develop little heat if operated within their ratings, they are especially suitable for use in equipment requiring minimum temperature variation.

Rectifier Ratings

Vacuum-tube rectifiers are subject to limitations as to breakdown voltage and current-handling capability. Some types are rated in terms of the maximum r.m.s. voltage which should be applied to the rectifier plate, while others, particularly mercury-vapor types, are rated according to maximum inverse peak voltage — the peak voltage between plate and cathode during the time the tube is not conducting. In the circuits shown in Fig. 7-2, the inverse peak voltage across each rectifier is 1.4 times the r.m.s. value of the voltage delivered by the entire transformer secondary.

The maximum d.c. output current is the maximum load current that can be drawn safely from the output of the filter. The value listed in tube tables is the safe maximum under average conditions. The exact value is dependent upon the nature of the filter that follows the rectifier.

A more significant rating is the maximum peak plate current. It is the peak value of the current pulses passing through the rectifier. This peak value can be much greater than the load current, especially if a large condenser is placed across the output of the rectifier because of the large instantaneous charging current drawn by the condenser if there is no impedance between the rectifier and the condenser. These peaks do not run as high with high-vacuum-type rectifiers as they do with rectifiers of the mercury-vapor type because of the relatively high series resistance of the former.

Rectifiers may be connected in parallel for current higher than the rated current of a single unit. This includes the use of the sections of a double diode for this purpose. Equalizing resistors of 50 to 100 ohms should be connected in series with each plate, as shown in Fig. 7-3, as a measure toward maintaining an equal division of current.

Operation of Rectifiers

In operating rectifiers requiring filament or cathode heating, care should be taken to provide the correct filament voltage at the tube terminals. Low filament voltage can cause excessive voltage drop in high-vacuum rectifiers and a considerable reduction in the inverse peak-voltage rating of a mercury-vapor tube. Filament connections to the rectifier socket should be firmly soldered, particularly in the case of the larger mercury-vapor tubes whose filaments operate at low voltage and high current. The socket should be selected with care, not only as to contact surface but also as to insulation, since the filament usually is at full output voltage to ground. Bakelite sockets will serve at voltages up to 500 or so, but ceramic sockets, well spaced from the chassis, always should be used at the higher voltages. Special filament transformers with high-voltage insulation between primary and secondary are required for rectifiers operating at potentials in excess of 1000 volts inverse peak.

The rectifier tubes should be placed in the equipment with adequate free space surrounding them to provide for proper ventilation. When mercury-vapor tubes are first placed in service, they should be allowed to run only with filament voltage for ten minutes before applying high voltage. After that, a delay of 30 seconds is recommended each time the filament is turned on.

Filters

The pulsating d.c. wave shown in Fig. 7-2 is not sufficiently smooth to prevent modulation. A filter consisting of chokes and condensers, as shown in Fig. 7-4, is connected between the rectifier output and the load circuit (transmitter or receiver) to smooth out the wave to the required degree.

The filter makes use of the energy-storage properties of the inductance of the choke and the capacitance of the condenser, energy being stored over the period during which the voltage and current are rising and releasing it to the load circuit during the period when the amplitude of the pulse is falling, thus leveling off the output by both lopping off the peaks and filling in the valleys.

Ripple Frequency and Voltage

The pulsations in the output of the rectifier can be considered to be the resultant of an alternating current superimposed upon a steady direct current. From this viewpoint, the filter may be considered to consist of shunting condensers which short-circuit the a.c. component while not interfering with the flow of the d.c. component, and series chokes which pass d.c. readily but which impede the flow of the a.c. component.

The alternating component is called the ripple. The effectiveness of the filter can be expressed in terms of per cent ripple which is the ratio of the r.m.s. value of the ripple to the d.c. value in terms of percentage. For c.w. transmitters, a reduction of the ripple to 5 per cent is considered adequate. The ripple in the output of power supplies for voice transmitters and VFOs should be reduced to 0.25 per cent or less. High-gain speech amplifiers and receivers may require a reduction to as low as 0.1 per cent to avoid objectionable hum.

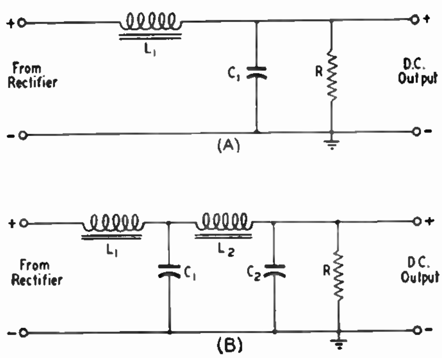


Fig. 7-4 — Choke-input filter circuits. A — Single-section. B — Double-section.

Ripple frequency is the frequency of the pulsations in the rectifier output wave — the number of pulsations per second. The frequency of the ripple with half-wave rectifiers is the same as the frequency of the line supply — 60 cycles with 60-cycle supply. Since the output pulses are doubled with a full-wave rectifier, the ripple frequency is doubled — to 120 cycles with 60-cycle supply.

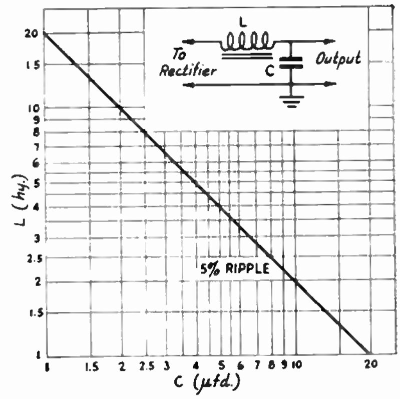


Fig. 7-5 — Graph showing combinations of inductance and capacitance that may be used to reduce ripple to 5 per cent with a single-section choke-input filter.

The amount of filtering (values of inductance and capacitance) required to give adequate smoothing depends upon the ripple frequency, more filtering being required as the ripple frequency is lower.

CHOKE-INPUT FILTERS

The filters shown in Fig. 7-4 are known as choke-input filters because the first element in the filter is a choke. This term is used in contrast to a condenser-input filter in which the first element is a condenser.

The percentage ripple output from a single-section filter (Fig. 7-4A) may be determined to a close approximation, for a ripple frequency of 120 cycles, from the following formula:

$$\left.\begin{array}{l}\text{Single-}\\\text{Section}\\\text{Filter}\end{array}\right\}\ \textit{Percentage ripple} = \frac{100}{LC}$$

where L is in hy. and C in μfd.

Example: L = 5 hy., C = 4 μfd.

$$\text{Percentage ripple} = \frac{100}{(5)\ (4)} = \frac{100}{20} = 5 \text{ per cent.}$$

Fig. 7-5 shows various other combinations of inductance and capacitance which will reduce the ripple to 5 per cent — the required minimum reduction for a supply for a c.w. transmitter.

Example: With a 10-hy. choke, what capacitance is required to reduce the ripple to 5 per cent?

Referring to Fig. 7-5, following the 10-hy. line horizontally, it intersects the ripple line at the 2-μfd. vertical line. Therefore the filter capacitance should be 2 μfd.

Example: With a 4-μfd. condenser, what choke inductance is required to reduce the ripple to 5 per cent?

Follow the vertical C = 4-μfd. line to the point where it intersects the ripple line; then follow the horizontal line at that point to read 5 hy., the required inductance.

In the case of a half-wave rectifier, the values of inductance and capacitance in the filter arrived at on the basis of a ripple frequency of 120 cycles must be doubled. It requires twice as much inductance and capacitance for the same degree of filtering with the half-wave circuit.

From the consideration of ripple reduction, any combination of inductances and capacitances which will give the required product and sum respectively will give the same ripple reduction. However, two other factors must be taken into consideration in the design of the filter. These are the peak rectifier current and voltage regulation.

Voltage Regulation

Unless the power supply is designed to prevent it, there may be a considerable difference between the output-terminal voltage of the supply when it is running free without an external load and the value when the external load is connected. Application of the load usually will be accompanied by a reduction in terminal voltage and this must be taken into consideration in the design of the supply. Regulation is commonly expressed as the percentage change in output voltage between no-load and full-load conditions in relation to the full-load voltage.

$$\textit{Per cent regulation} = \frac{100\,(E_1 - E_2)}{E_2}.$$

Example: No-load voltage = E_1 = 1550 volts.
Full-load voltage = E_2 = 1230 volts.

$$\text{Percentage regulation} = \frac{100\,(1550 - 1230)}{1230} = \frac{32{,}000}{1230} = 26 \text{ per cent.}$$

With proper design and the use of conservatively-rated components, a regulation of 10 per cent or less at the output terminals of the supply unit is possible with a choke-input filter. Good voltage regulation may or may not be of primary importance depending upon the nature of the load. If the load is constant, as in the case of a receiver, speech amplifier or the stages of a transmitter which are not keyed, voltage regulation, so far as that contributed by filter design is concerned, may be of secondary importance. The highly-stabilized voltage desirable for high frequency-stability of oscillators in receivers and transmitters is obtained by other means. Power supplies for the keyed stage of a c.w. transmitter and the stages following, and for Class B modulators, should have good regulation.

Bleeder Resistor

In general, a bleeder resistor is a resistance connected across the output of a filter to supply a minimum load (see R, Fig. 7-4). It also serves as a safety measure to discharge the filter condensers when the supply is turned off. The bleeder resistance need not be composed entirely of a resistor. Any constant load on the supply may serve the same purpose. In this case, a resistor of a high value should be used as a protective device to discharge the filter condensers.

The Input Choke

The rectifier peak current and the power-supply voltage regulation depend almost entirely upon the inductance of the input choke in relation to the load resistance. The function of the choke is to raise the ratio of average to peak current (by its energy storage), and to prevent the d.c. output voltage from rising above the average value of the a.c. voltage applied to the rectifier. For both purposes, its impedance to the flow of the a.c. component must be high.

The value of input-choke inductance which prevents the d.c. output voltage from rising above the average of the rectified a.c. wave is the **critical inductance**. For 120-cycle ripple, it is given by the approximate formula:

$$L_{\text{crit.}} = \frac{\textit{Load resistance}\text{ (ohms)}}{1000}.$$

For other ripple frequencies, the inductance required will be the above value multiplied by the ratio of 120 to the actual ripple frequency.

With inductance values less than critical, the d.c. output voltage will rise because the filter tends to act as a condenser-input filter. With critical inductance, the peak plate current of one tube in a center-tap rectifier will be approximately 10 per cent higher than the d.c. load current taken from the supply.

An inductance of twice the critical value is called the **optimum value**. This value gives a further reduction in the ratio of peak-to-average plate current, and represents the point at which further increase in inductance does not give correspondingly improved operating characteristics.

Swinging Chokes

The formula for critical inductance indicates that the minimum inductance required varies widely with the load resistance. In the case where there is no load except the bleeder on the power supply, the critical inductance required is the highest: much lower values are satisfactory when the full-load current is being delivered. Since the inductance of a choke tends to rise as the direct current flowing through it is decreased, it is possible to effect an economy in materials by designing the choke to have a "swinging" characteristic so that it has the required critical inductance value with the bleeder load only, and about the optimum inductance value at full load. If the bleeder resistance is 20,000 ohms and the full-load resistance (including the bleeder) is 2500 ohms, a choke which swings from 20 henrys to 5 henrys over the full output-current range will fulfill the requirements. With any given input choke, the bleeder resistance (or other steady minimum load) should be 1000 times the maximum inductance of the choke in henrys.

Example: With a swinging choke of 5 to 20 hy., the bleeder resistance (or the resultant of the bleeder plus other steady load in parallel) should not exceed (20) (1000) = 20,000 ohms.

Output Condenser

If the supply is intended for use with an audio-frequency amplifier, the reactance of the last filter condenser should be small (20 per cent or less) compared with the other a.f. resistance or impedance in the circuit, usually the tube plate resistance and load resistance. On the basis of a lower a.f. limit of 100 cycles for speech amplification, this condition usually is satisfied when the output capacitance (last filter capacitor) of the filter is 4 to 8 μfd., the higher value of capacitance being used in the case of lower tube and load resistances.

Resonance

Resonance effects in the series circuit across the output of the rectifier which is formed by the first choke (L_1) and first filter condenser (C_1) must be avoided, since the ripple voltage would build up to large values. This not only is the opposite action to that for which the filter is intended, but also may cause excessive rectifier peak currents and abnormally-high inverse peak voltages. For full-wave rectification the ripple frequency will be 120 cycles for a 60-cycle supply, and resonance will occur

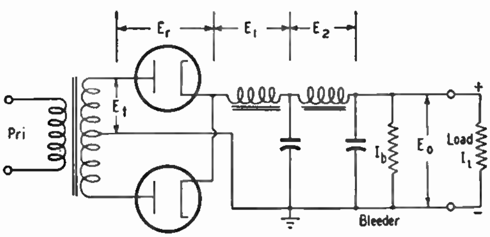


Fig. 7-6 — Diagram showing various voltage drops that must be taken into consideration in determining the required transformer voltage to deliver the desired output voltage.

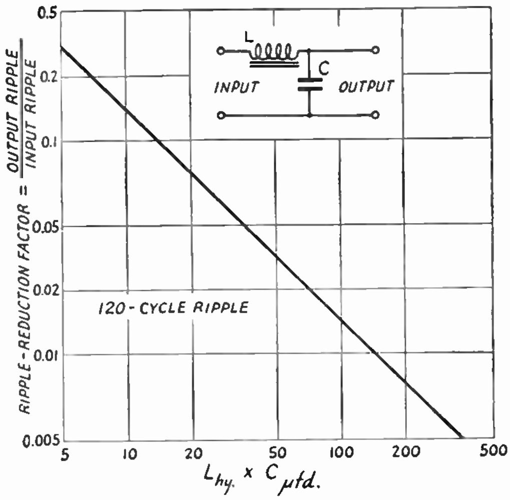


Fig. 7-7 — Ripple-reduction factor for various values of *L* and *C* in filter section. *Output ripple = input ripple × ripple factor.*

when the product of choke inductance in henrys times condenser capacitance in microfarads is equal to 1.77. The corresponding figure for 50-cycle supply (100-cycle ripple frequency) is 2.53, and for 25-cycle supply (50-cycle ripple frequency) 13.5. At least twice these products should be used to ensure against resonance effects.

Output Voltage

Provided the input-choke inductance is at least the critical value, the output voltage may be calculated quite closely by the following equation:

$$E_o = 0.9E_t - \frac{(I_b + I_L)(R_1 + R_2)}{1000} - E_r$$

where E_o is the output voltage; E_t is the r.m.s. voltage applied to the rectifier (r.m.s. voltage between center-tap and one end of the secondary in the case of the center-tap rectifier); I_b and I_L are the bleeder and load currents, respectively, in milliamperes; R_1 and R_2 are the resistances of the first and second filter chokes; and E_r is the drop between rectifier plate and cathode. These voltage drops are shown in Fig. 7-6. At no load I_L is zero, hence the no-load voltage may be calculated on the basis of bleeder current only. The voltage regulation may be determined from the no-load and full-load voltages.

Additional Filtering

The graph of Fig. 7-7 shows the factor by which the ripple percentage may be reduced by the addition of one or more sections of filter, each similar in configuration to the first.

Example:

Ripple after first section = 5 per cent.
L in second section = 10 hy.
C in second section = 8 μfd.
L × *C* = 80.

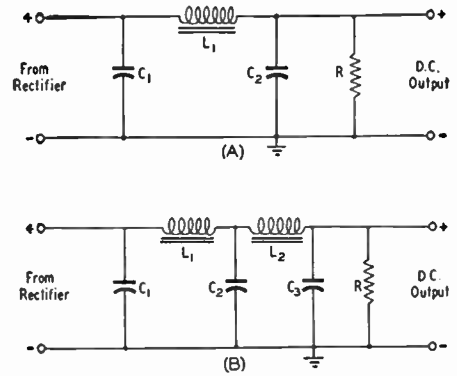


Fig. 7-8 — Condenser-input filter circuits. A — Single-section. B — Double-section.

From Fig. 7-7, the reduction factor is approximately 0.019. Therefore the ripple after the second section will be 5 × 0.019 = 0.095 per cent.

CONDENSER-INPUT FILTERS

Condenser-input filters are shown in Fig. 7-8. In comparison with a properly-designed input-choke filter, the d.c. output voltage is higher for most values of load, the ratio of peak rectifier plate current to d.c. output current is greater and the voltage regulation is considerably poorer.

The approximate performance of a filter consisting of the input condenser only is indicated in Figs. 7-9, 7-10 and 7-11. Fig. 7-9 shows the relationship between rectifier peak plate current and d.c. load current for various values of load and input resistance. Input resistance is the sum of transformer and rectifier resistances. In each case a capacitance of 4 μfd. or greater is assumed, since the characteristics change relatively little with higher values of capacitance.

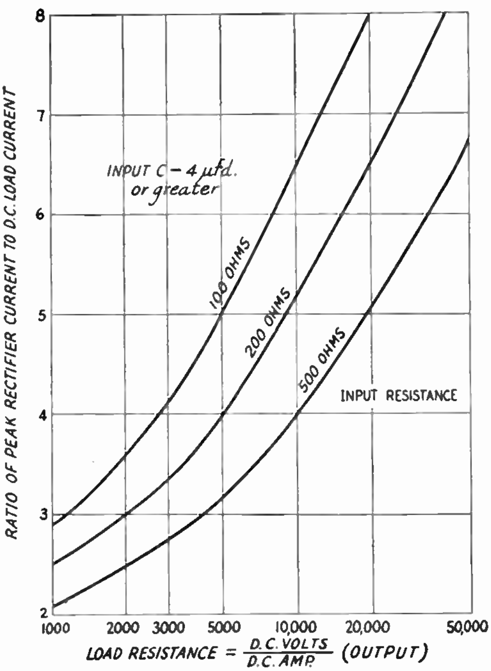


Fig. 7-9 — Graph showing relationship between d.c. load current and rectifier peak plate current with condenser input for various load and input resistances.

Fig. 7-10 shows the ratio of d.c. output voltage to the transformer r.m.s. voltage. In this respect, too, the change with higher capacitance values is small.

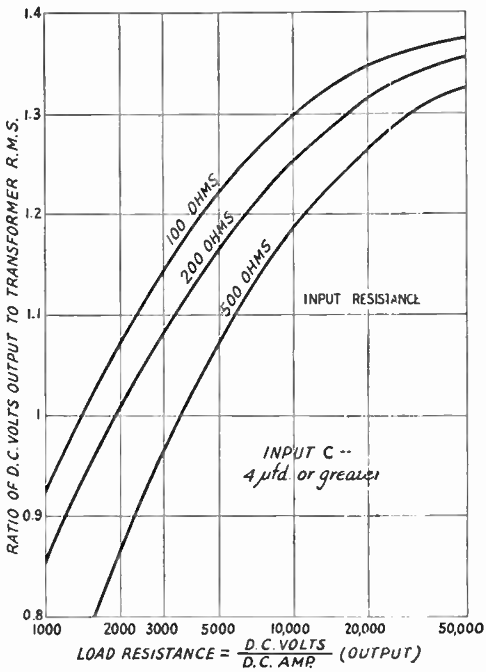


Fig. 7-10 — Chart showing approximate ratio of d.c. output voltage across filter input condenser to transformer r.m.s. secondary voltage for different load and input resistances.

Fig. 7-11 shows the approximate percentage ripple across the input condenser for capacitances of 4 and 8 μfd. The change in ripple voltage with normal differences in input resistance is relatively slight.

Further reduction in ripple may be obtained by adding sections of series inductance and parallel capacitance, as shown in Fig. 7-8. The reduction factor from Fig. 7-7 applies in this case also.

Example:

Input condenser — 4 μfd.
Output condenser — 8 μfd.
Input resistance — 200 ohms.
Transformer r.m.s. voltage — 400.
Load resistance (including resistance of filter choke) — 5000 ohms.

From Fig. 7-10, $\frac{\text{D.c. volts output}}{\text{Transformer r.m.s.}} = 1.17.$

D.c. volts output = 400 × 1.17 = 468 volts.

From Fig. 7-9, $\frac{\text{Peak rectifier current}}{\text{D.c. load current}} = 4.$

D.c. load current $= \frac{468}{5000} = 93.6$ ma.

Peak rectifier current $= 93.5 \times 4 = 374$ ma.

From Fig. 7-11, ripple percentage across input condenser = approximately 8 per cent.

$L \times C = 8 \times 20 = 160.$

From Fig. 7-7, reduction factor = 0.009.

Output ripple percentage $= 8 \times 0.009 = 0.072$ per cent.

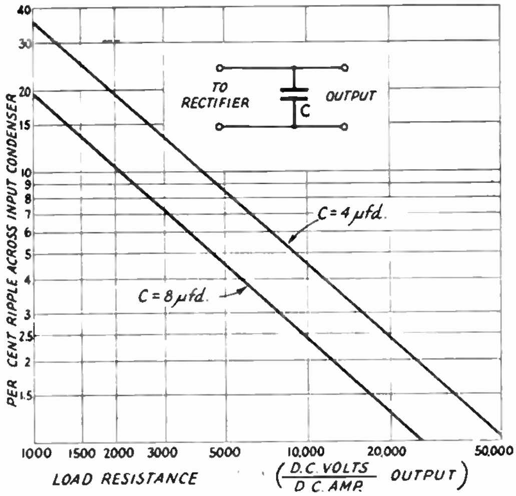


Fig. 7-11 — Chart showing approximate 120-cycle percentage ripple across filter input condenser for various loads.

RATINGS OF FILTER COMPONENTS

Although filter condensers in a choke-input filter are subjected to smaller variations in d.c. voltage than in the condenser-input filter, it is advisable to use condensers rated for the peak transformer voltage in case the bleeder resistor should burn out when there is no load on the power supply, since the voltage then will rise to the same maximum value as with a condenser-input filter.

In a condenser-input filter, the condensers should have a working-voltage rating at least as high and preferably somewhat higher, as a safety factor. Thus, in the case of a center-tap rectifier having a transformer delivering 550 volts each side of the center-tap, the minimum safe condenser voltage rating will be 550 × 1.41 or 775 volts. An 800-volt condenser should be used, or preferably a 1000-volt unit to allow a margin of safety.

Filter condensers are made in several different types. Electrolytic condensers, which are available for voltages up to about 800, combine high capacitance with small size, since the dielectric is an extremely-thin film of oxide on aluminum foil. Condensers for higher voltages usually are made with a dielectric of thin paper impregnated with oil. The **working voltage** of a condenser is the voltage that it will withstand continuously.

The input choke may be of the swinging type, the required no-load and full-load inductance values being calculated as described above. The second choke **(smoothing choke)** should have constant inductance with varying d.c. load currents. Values of 10 to 20 henrys ordinarily are used. Since chokes usually are placed in the positive leads, the negative being grounded, the windings should be insulated from the core to withstand the full d.c. output voltage of the supply and be capable of handling the required load current.

Filter chokes or inductances are wound on iron cores, with a small gap in the core to prevent magnetic saturation of the iron at high currents. When the iron becomes saturated its permeability decreases, consequently the inductance also decreases. Despite the air gap, the inductance of a choke usually varies to some extent with the direct current flowing in the winding; hence it is necessary to specify the inductance at the current which the choke is intended to carry. Its inductance with little or no direct current flowing in the winding may be considerably higher than the value when full load current is flowing.

The Plate Transformer

Output Voltage

The output voltage which the plate transformer must deliver depends upon the required d.c. load voltage and the type of rectifier circuit. With condenser-input filters, the r.m.s. secondary voltage usually is made equal to or slightly more than the d.c. output voltage, allowing for voltage drops in the rectifier tubes and filter chokes as well as in the transformer itself. The full-wave center-tap rectifier requires a transformer giving this voltage each side of the secondary center-tap, the *total* secondary voltage being twice the desired d.c. output voltage.

With a choke-input filter, the required r.m.s. secondary voltage (each side of center-tap for a center-tap rectifier) can be calculated by the equation:

$$E_t = 1.1\left[E_o + \frac{I(R_1 + R_2)}{1000} + E_r\right]$$

where E_o is the required d.c. output voltage, I is the load current (including bleeder current) in ma., R_1 and R_2 are the resistances of the chokes, and E_r is the voltage drop in the rectifier. E_t is the full-load r.m.s. secondary voltage; the open-circuit voltage usually will be 5 to 10 per cent higher than the full-load value.

Volt-Ampere Rating

The volt-ampere rating of the transformer depends upon the type of filter (condenser or choke input). With a condenser-input filter the heating effect in the secondary is higher because of the high ratio of peak to average current, consequently the volt-amperes consumed by the transformer may be several times the watts delivered to the load. With a choke-input filter, provided the input choke has at least the critical inductance, the secondary volt-amperes can be calculated quite closely by the equation:

$$Sec.\ V.A. = 0.00075EI$$

where E is the *total* r.m.s. voltage of the secondary (between the outside ends in the case of a center-tapped winding) and I is the d.c. output current in milliamperes (load current plus bleeder current). The primary volt-amperes will be 10 to 20 per cent higher because of transformer losses.

Building Small Transformers

Power transformers for both filament heating and plate supply for all transmitting and rectifying tubes are available commercially, but occasionally the amateur wishes to build a transformer for some special purpose or has a core from a burned-out transformer on which he wishes to put new windings.

Most transformers that amateurs build are for use on 115-volt 60-cycle supplies. The number of turns necessary on the 115-volt winding depends on the kind of iron used in the core and on the cross-sectional area of the core. Silicon steel is best, and a flux density of about 50,000 lines per square inch can be used. This is the basis of the table of cross-sections given.

An average value for the number of primary turns to be used is 7.5 turns per volt per square inch of cross-sectional area. This relation may be expressed as follows:

$$No.\ primary\ turns = 7.5\left(\frac{E}{A}\right)$$

where E is the primary voltage and A the number of square inches of cross-sectional area of the core. For 115-volt primary transformers the equation becomes:

$$No.\ primary\ turns = \frac{863}{A}.$$

When a small transformer is built to handle a continuous load, the copper wire in the windings should have an area of 1500 circular mils for each ampere carried. (See Wire Table in Chapter Twenty-Four.) For intermittent use, 1000 circular mils per ampere is permissible.

The primary wire size is given in Table 7-I; the secondary wire size should be chosen according to the current to be carried, as previously described. The Wire Table in Chapter Twenty-Four shows how many turns of each wire size can be wound into a square inch of window area, assuming that the turns are wound regularly and that no insulation is used between layers. The primary winding of a 200-watt transformer, which has 270 turns of No. 17 wire, would occupy 270/329 or 0.82 square inch if wound with double-cotton-covered wire, for example. This makes no allowance for a layer of insulation between the windings (in general, it is good practice to wind a strip of paper between each layer) so that the winding-area allowance should be increased if layer insulation is to be used. The figures also are based on accurate winding such as is done by machines; with hand-winding it is probable that somewhat more area would be required. An increase of 50 per cent should take care of both hand winding and layer thickness. The area to be taken by the secondary winding should be estimated, as should also the area likely to be occupied by the insulation between the core and windings and between the primary and secondary windings themselves. When the total window area required has been figured —

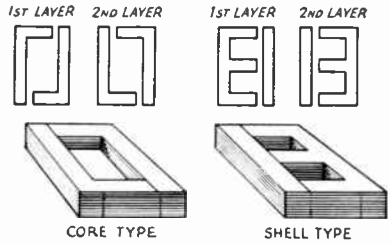


Fig. 7-12 — Two different types of transformer cores and their laminations.

TABLE 7-I
Transformer Design

Input (Watts)	*Full-Load Efficiency*	*Size of Primary Wire*	*No. of Primary Turns*	*Turns Per Volt*	*Cross-Section Through Core*
50	75%	23	528	4.80	1¼" × 1¼"
75	85	21	437	3.95	1⅜ × 1⅜
100	90	20	367	3.33	1½ × 1½
150	90	18	313	2.84	1⅝ × 1⅝
200	90	17	270	2.45	1¾ × 1¾
250	90	16	248	2.25	1⅞ × 1⅞
300	90	15	248	2.25	1⅞ × 1⅞
400	90	14	206	1.87	2 × 2
500	95	13	183	1.66	2⅛ × 2⅛
750	95	11	146	1.33	2⅜ × 2⅜
1000	95	10	132	1.20	2½ × 2½
1500	95	9	109	0.99	2¾ × 2¾

allowing a little extra for contingencies — laminations having the desired leg-width and window area should be purchased. It may not be possible to get laminations having exactly the dimensions wanted, in which case the nearest size should be chosen. The cross-section of the core need not be square but can be rectangular in shape so long as the core area is great enough. It is easier to wind coils for a core of square cross-section, however.

Transformer cores are of two types, "core" and "shell." In the core type, the core is simply a hollow rectangle formed from two "L"-shaped laminations, as shown in Fig. 7-12. Shell-type laminations are "E"- and "I"-shaped, the transformer windings being placed on the center leg. Since the magnetic path divides between the outer legs of the "E," these legs are each half the width of the center leg. The cross-sectional area of a shell-type core is the cross-sectional area of the center leg. The shell-type core makes a better transformer than the core type, because it tends to prevent leakage of the magnetic flux. Calculations are the same for both types.

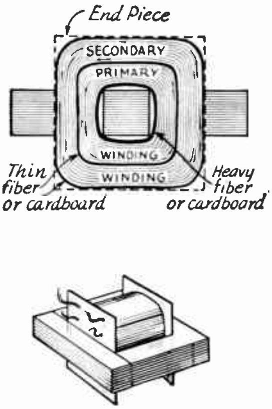


Fig. 7-13 — A convenient method of assembling the windings of a shell-type core. Windings can be similarly mounted on core-type cores, in which case the coils are placed on one of the sides. High-voltage core-type transformers sometimes are made with the primary on one core leg and the secondary on the opposite.

Fig. 7-13 shows the method of putting the windings on a shell-type core. The primary is usually wound on the inside — next to the core — on a form made of fiber or several layers of cardboard. This form should be slightly larger than the core leg on which it is to fit so that it will be an easy matter to slip in the laminations after the coils are completed and ready for mounting. The terminals are brought out to the side. After the primary is finished, the secondary is wound over it, several layers of insulating material being put between. If the transformer is for high voltages, the high-voltage winding should be carefully insulated from the primary and core by a few layers of Empire Cloth or tape. A protective covering of heavy cardboard or thin fiber should be put over the outside of the secondary to protect it from damage and to prevent the core from rubbing through the insulation. Square-shaped end pieces of fiber or cardboard usually are provided to protect the sides of the windings and to hold the terminal leads in place. High-voltage terminal leads should be enclosed either in Empire Cloth tubing or in spaghetti.

After the windings are finished the core should be inserted, one lamination at a time. Fig. 7-12 shows the method of building up the core. Alternate "E"-shaped laminations are pushed through the core opening from opposite sides. The "I"-shaped laminations are used to fill the end spaces, butting against the open ends of the "E"-shaped pieces. This method of building up the core ensures a good magnetic path of low reluctance. All laminations should be insulated from each other to prevent eddy currents from flowing. If there is iron rust or scale on the core material, that will serve the purpose very well — otherwise one side of each piece can be coated with thin shellac. It is essential that the joints in the core be well-made and be square and even. After the transformer is assembled, the joints can be hammered up tight using a block of wood between the hammer and the core to prevent damaging the laminations. If the winding form does not fit tightly on the core, small wooden wedges may be driven between it and the core to prevent vibration. Transformers built by the amateur can be painted with insulating varnish or waxed to make them rigid and moistureproof. A solution of melted beeswax and rosin makes a good impregnating mixture. Melted paraffin should not be used because it has too low a melting point. Double-cotton-covered wire can be coated with shellac as each layer is put on. However, enameled wire should never be treated with shellac as it may dissolve the enamel and hurt the insulation, and it will not dry because the moisture in the shellac will not be absorbed by the insulation. Small transformers can be treated with battery compound after they are wound and assembled. Strips of thin paper between layers of small enameled wire are necessary to keep each layer even and to give added insulation. Thick paper must be avoided since it keeps in the heat generated in the winding so that the temperature may become dangerously high.

Keep watch for shorted turns and layers. If just a single turn should become shorted in the entire winding, the voltage set up in it would cause a heavy current to flow which would burn it up, making the whole transformer useless.

Taps can be taken off as the windings are made if it is desired to have a transformer giving several voltages. Taps should be arranged, whenever possible, so that they come at the ends of the layers.

After leaving the transformer primary winding connected to the line for several hours it should be only slightly warm. If it draws much current or gets excessively hot there is something wrong. Some short-circuited turns are probably responsible and will continue to cause overheating.

Voltage Dropping

Series Voltage-Dropping Resistor

Certain plates and screens of the various tubes in a transmitter or receiver often require a variety of operating voltages differing from the output voltage of available power supplies. In most cases, it is not economically feasible to provide a separate power supply for each of the required voltages. If the current drawn by an electrode, or combination of electrodes operating at the same voltage, is reasonably constant under normal operating conditions, the required voltage may be obtained from a supply of higher voltage by means of a voltage-dropping resistor in series, as shown in Fig. 7-14A. The value of the series resistor, R_1, may be obtained from Ohm's Law, $R = \frac{E_d}{I}$, where E_d is the voltage *drop* required from the supply voltage to the desired voltage and I is the total rated current of the load.

Example: The plate of the tube in one stage and the screens of the tubes in two other stages require an operating voltage of 250. The nearest available supply voltage is 400 and the total of the rated plate and screen currents is 75 ma. The required resistance is

$$R = \frac{400 - 250}{0.075} = \frac{150}{0.075} = 2000 \text{ ohms.}$$

The power rating of the resistor is obtained from P (watts) $= I^2R = (0.075)^2\,(2000) = 11.2$ watts. A 25-watt resistor is the nearest safe rating to be used.

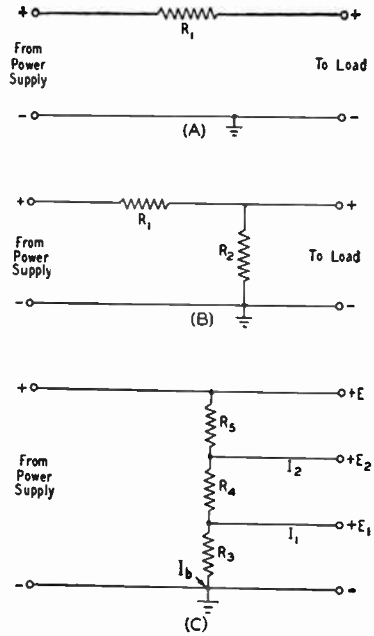


Fig. 7-14 — A — Series voltage-dropping resistor. B — Simple voltage divider. C — Multiple divider circuit.

$$R_3 = \frac{E_1}{I_b};\ R_4 = \frac{E_2 - E_1}{I_b + I_1};\ R_5 = \frac{E - E_2}{I_b + I_1 + I_2}$$

Voltage Dividers

The regulation of the voltage obtained in this manner obviously is poor, since any change in current through the resistor will cause a directly-proportional change in the voltage drop across the resistor. The regulation can be improved somewhat by connecting a second resistor from the low-voltage end of the first to the negative power-supply terminal, as shown in Fig. 7-14B. Such an arrangement constitutes a **voltage divider.** The second resistor, R_2, acts as a constant load for the first, R_1, so that any variation in current from the tap becomes a smaller percentage of the total current through R_1. The heavier the current drawn by the resistors when they alone are connected across the supply, the better will be the voltage regulation at the tap.

Such a voltage divider may have more than a single tap for the purpose of obtaining more than one value of voltage. A typical arrangement is shown in Fig. 7-14C. The terminal voltage is E, and two taps are provided to give lower voltages, E_1 and E_2, at currents I_1 and I_2 respectively. The smaller the resistance between taps in proportion to the total resistance, the smaller the voltage between the taps. For convenience, the voltage divider in the figure is considered to be made up of separate resistances R_3, R_4, R_5, between taps. R_3 carries only the bleeder current, I_b; R_4 carries I_1 in addition to I_b; R_5 carries I_2, I_1 and I_b. To calculate the resistances required, a bleeder current, I_b, must be assumed; generally it is low compared with the total load current (10 per cent or so). Then the required values can be calculated as shown in Fig. 7-14C, I being in amperes.

The method may be extended to any desired number of taps, each resistance section being calculated by Ohm's Law using the voltage drop across it and the total current through it. The power dissipated by each section may be calculated either by multiplying I and E or I^2 and R.

Voltage Stabilization

Gaseous Regulator Tubes

There is frequent need for maintaining the voltage applied to a low-voltage low-current circuit at a practically constant value, regardless of the voltage regulation of the power supply or variations in load current. In such applications, gaseous regulator tubes (VR105-30, VR150-30, etc.) can be used to good advantage. The voltage drop across such tubes is constant over a moderately wide current range. Tubes are available for regulated voltages of 150, 105, 90 and 75 volts.

The fundamental circuit for a gaseous regulator is shown in Fig. 7-15A. The tube is connected in series with a **limiting resistor,** R_1, across a source of voltage that must be higher than the **starting** voltage. The starting voltage is about 30 per cent higher than the operating voltage. The load is connected in parallel with the tube. For stable operation, a minimum tube current of 5 to 10 ma. is required. The maximum permissible current with most types is 40 ma.; consequently, the load current cannot exceed 30 to 35 ma. if the voltage is to be stabilized over a range from zero to maximum load current.

The value of the limiting resistor must lie between that which just permits minimum tube current to flow and that which just passes the maximum permissible tube current when there is no load current. The latter value is generally used. It is given by the equation:

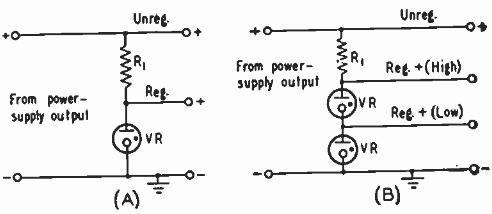


Fig. 7-15 — **Voltage-stabilizing circuits using VR tubes.**

$$R = \frac{1000\,(E_s - E_r)}{I}$$

where R is the limiting resistance in ohms, E_s is the voltage of the source across which the tube and resistor are connected, E_r is the rated voltage drop across the regulator tube, and I is the maximum tube current in milliamperes (usually 40 ma.).

Fig. 7-15B shows how two tubes may be used in series to give a higher regulated voltage than is obtainable with one, and also to give two values of regulated voltage. The limiting resistor may be calculated as above, using the sum of the voltage drops across the two tubes for E_r. Since the upper tube must carry more current than the lower, the load connected to the low-voltage tap must take small current. The total current taken by the loads on both the high and low taps should not exceed 30 to 35 milliamperes.

Voltage regulation of the order of 1 per cent can be obtained with regulator circuits of this type.

Electronic Voltage Regulation

A voltage-regulator circuit suitable for higher voltages and currents than the gaseous tubes, and also having the feature that the output voltage can be varied over a rather wide range, is shown in Fig. 7-16. A high-gain voltage-amplifier tube, usually a sharp cut-off pentode, is connected in such a way that a small change in the output voltage of the power supply causes a change in grid bias, and thereby a corresponding change in plate current. Its plate current flows through a resistor (R_5), the voltage drop across which is used to bias a second tube — the "regulator" tube — whose plate-cathode circuit is connected in series with the load circuit. The regulator tube therefore functions as an automatically-variable series resistor. Should the output voltage increase slightly the bias on the control tube will become more positive, causing the plate current of the control tube to increase and the drop across R_5 to increase correspondingly. The bias on the regulator tube therefore becomes more negative and the effective resistance of the regulator tube increases, causing the terminal voltage to drop. A decrease in output voltage causes the reverse action. The time lag in the action of the system is negligible, and with proper circuit constants the output voltage can be held within a fraction of a per cent throughout the useful range of load current and over a wide range of supply.

An essential in this system is the use of a constant-voltage bias source for the control tube. The voltage change which appears at the grid of the tube is the *difference* between a fixed negative bias and a positive voltage which is taken from the voltage divider across the output. To get the most effective control, the negative bias must not vary with plate current. The most satisfactory type of bias is a dry battery of 45 to 90 volts, but a gaseous regulator tube (VR75-30) or a neon bulb of the type without a resistor in the base may be used instead. If the gas tube or neon bulb is used, a negative-resistance type of oscillation may take place at audio frequencies or higher, in which case a condenser of 0.1 μfd. or more should be connected across the tube. A similar condenser between the control-tube grid and cathode also is frequently helpful in this respect.

The variable resistor, R_3, is used to adjust

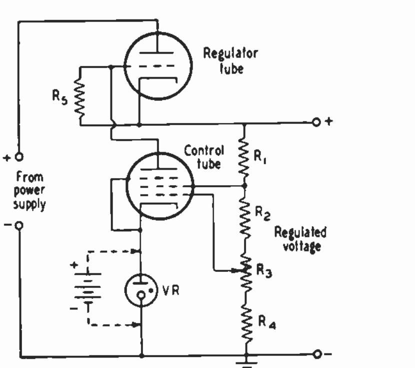


Fig. 7-16 — **Electronic voltage regulator. The regulator tube is ordinarily a 2A3 or a number of them in parallel, the control tube a 6SJ7 or similar type. The filament transformer for the regulator tube must be insulated for the plate voltage, and cannot supply current to other tubes when a filament-type regulator tube is used. Typical values: R_1, 10,000 ohms; R_2, 22,000 ohms; R_3, 10,000-ohm potentiometer; R_4, 4700 ohms; R_5, 0.47 megohm.**

Fig. 7-17 — A heavy-duty electronically-regulated power supply. The unit is assembled on a 6 × 14 × 3-inch chassis fitted with an enclosing cover. The five tubes across the rear, left to right, are the 6AS7G regulator, the 6SJ7 control tube, the VR-105 bias regulator, the 1-V bias rectifier and the 5U4G power rectifier. In the foreground are the two filter chokes and the power transformer. The remainder of the components are mounted underneath.

the bias on the control tube to the proper operating value. It also serves as an output voltage control, setting the value of regulated voltage within the existing operating limits.

The maximum output voltage obtainable is equal to the power-supply voltage minus the minimum drop through the regulator tube. This drop is of the order of 50 volts with the tubes ordinarily used. The maximum current also is limited by the regulator tube; 100 milliamperes is a safe value for the 2A3. Two or more regulator tubes may be connected in parallel to increase the current-carrying capacity, without need for changes in the circuit arrangement.

A heavy-duty regulated supply of this type is shown in Fig. 7-17. The circuit is shown in Fig. 7-18. A 6AS7G dual power triode is used as the regulator which is controlled by a 6SJ7. Reference bias is furnished by means of a 1-V half-wave rectifier whose output is regulated by a VR-105 regulator tube. The supply is capable of delivering 150 ma. over a range of 120 to 340 volts. Filament voltage and an external connection from the bias supply are also brought out to the output socket.

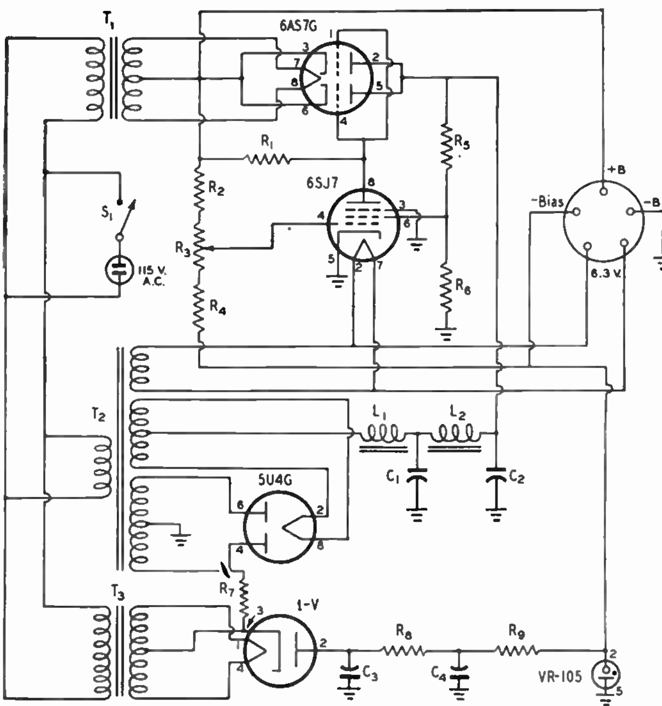


Fig. 7-18 — Circuit diagram of the electronically-regulated power supply.

C_1, C_2, C_3, C_4 — 16-μfd. 450-volt electrolytic.
R_1 — 0.47 megohm, ½ watt.
R_2 — 0.18 megohm, ½ watt.
R_3 — 75,000-ohm potentiometer.
R_4 — 0.1 megohm, ½ watt.
R_5 — 50,000 ohms, 10 watts.
R_6 — 24,000 ohms, 2 watts.
R_7, R_8, R_9 — 2500 ohms, 10 watts.
L_1 — 8/30-hy. 150-ma. filter choke (Stancor C1718).
L_2 — 30-hy. 110-ma. filter choke (Stancor C1001).
S_1 — S.p.s.t. toggle switch.
T_1 — Filament transformer: 6.3 volts, 3 amp. (Stancor P-5014).
T_2 — Power transformer: 375-0-375 volts, 150 ma.; 5 volts, 3 amp.; 6.3 volts, 5 amp. (Stancor P-6014).
T_3 — Filament transformer: 6.3 volts, 1.2 amp. (Stancor P-6134).

Miscellaneous Power-Supply Circuits

Duplex Plate Supplies

In some cases it may be advantageous economically to obtain two plate-supply voltages from a single power supply, making one or more of the components serve a double purpose. Circuits of this type are shown in Figs. 7-19 and 7-20.

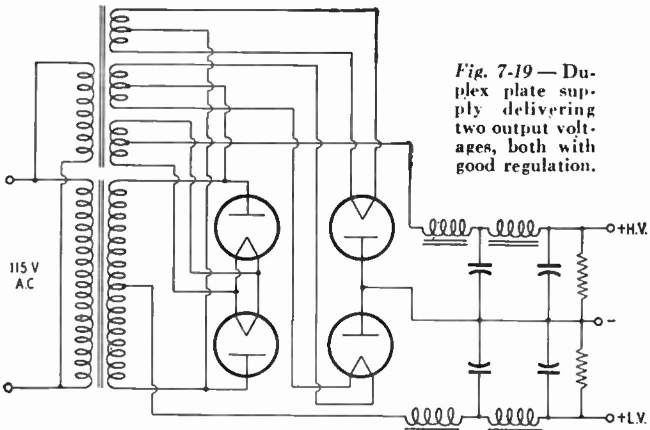


Fig. 7-19 — Duplex plate supply delivering two output voltages, both with good regulation.

In Fig. 7-19, a bridge rectifier is used to obtain the full transformer voltage, while a connection is also brought out from the center-tap to obtain a second voltage corresponding to half the total transformer secondary voltage. The sum of the currents drawn from the two taps should not exceed the d.c. ratings of the rectifier tubes and transformer. Filter values for each tap are computed separately.

Fig. 7-20 shows how a transformer with multiple secondary taps may be used to obtain both high and low voltages simultaneously. A separate full-wave rectifier is used at each pair of taps. The filter chokes are placed in the common negative lead, but separate filter condensers are required. The sum of the currents drawn from each pair of taps must not exceed the transformer rating, and the chokes must carry the total load current. Each bleeder should have a value in ohms 1000 times the maximum rated inductance in henrys of the swinging choke, L_1, for best regulation. A power supply of this type is shown in Figs. 7-21 and 7-22. In this case two sets of chokes are used to divide the load current.

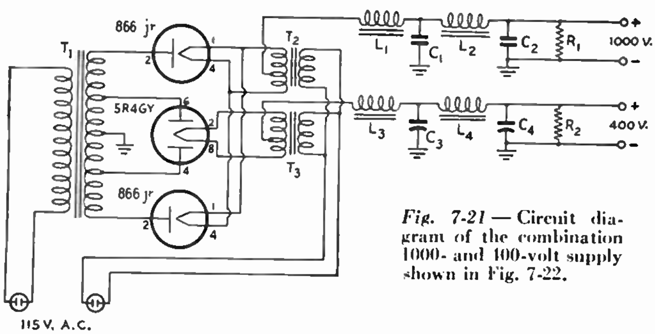


Fig. 7-21 — Circuit diagram of the combination 1000- and 400-volt supply shown in Fig. 7-22.

C_1, C_2 — 2-μfd. 1000-volt paper (Mallory TX805).
C_3 — 4-μfd. 600-volt electrolytic (C-D 604).
C_4 — 8-μfd. 600-volt electrolytic (C-D 608).
R_1 — 20,000 ohms, 75 watts.
R_2 — 20,000 ohms, 25 watts.
L_1, L_3 — 5/20-hy. swinging choke, 150 ma. (Thordarson T-19C39).
L_2, L_4 — 12-hy. smoothing choke. 150 ma. (Thordarson T-19C46).
T_1 — High-voltage transformer, 1075 and 500 volts r.m.s. each side, 125- and 150-ma. simultaneous current rating (Thordarson T-19P57).
T_2 — 2.5 volts, 5 amp. (Thordarson T-19F88).
T_3 — 5 volts, 4 amp. (Thordarson T-63F99).

Selenium-Rectifier Circuits

While the circuits shown in Figs. 7-23, 7-24 and 7-25 may be used with any type of rectifier, they find their greatest advantage when used with selenium rectifiers which require no filament transformer.

Fig. 7-23 is a straightforward half-wave rectifier circuit which may be used in applications where 115 to 130 volts d.c. is desired. It makes an ideal bias supply, for instance. In this, as well as other circuits, it will be observed that the negative side of the output is common with one side of the a.c. line and it is suggested that this side be fused with a ½-ampere fuse.

Fig. 7-24 shows several voltage-doubler circuits. Of the three, the one

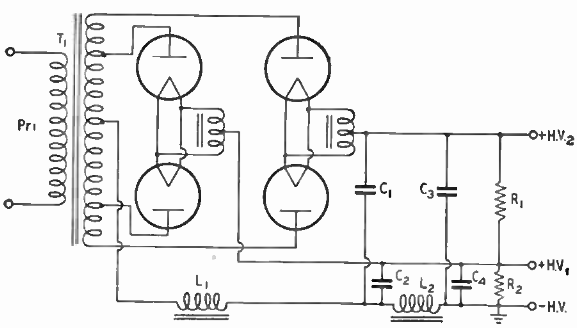
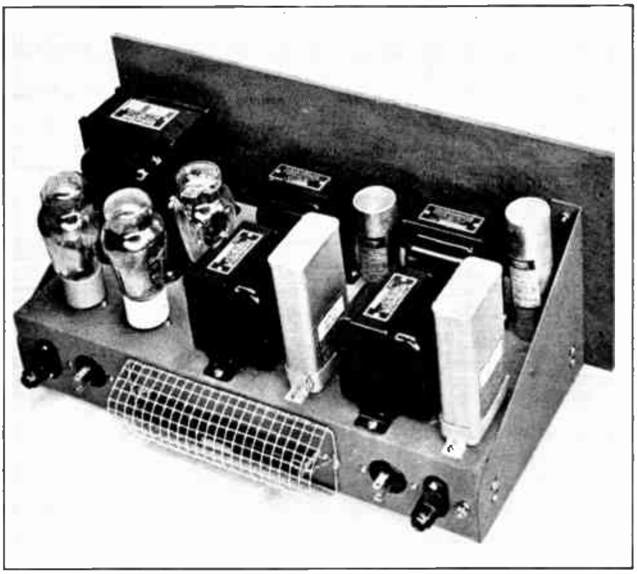


Fig. 7-20 — Power supply in which a single transformer and set of chokes serve for two different output voltages.

Fig. 7-22 — This power supply makes use of a combination transformer and a dual filter system, delivering 1000 volts at 125 ma. and 400 volts at 150 ma., or 400 volts and 750 volts simultaneously, depending upon the transformer selected. The circuit diagram is given in Fig. 7-21. The 1000-volt bleeder resistor is mounted on the rear edge of the chassis, with a protective guard made of a piece of galvanized fencing material to provide ventilation. Millen safety terminals are used for the two high-voltage terminals. Ceramic sockets should be used for the 866 Jrs. The chassis measures 8 by 17 by 3 inches.



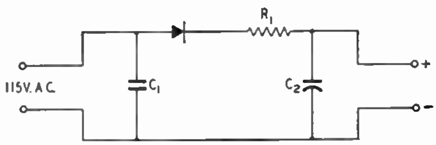


Fig. 7-23 — Simple half-wave circuit for selenium rectifier.

C_1 — 0.05-μfd. 600-volt paper.
C_2 — 40-μfd. 200-volt electrolytic.
R_1 — 25 to 100 ohms.

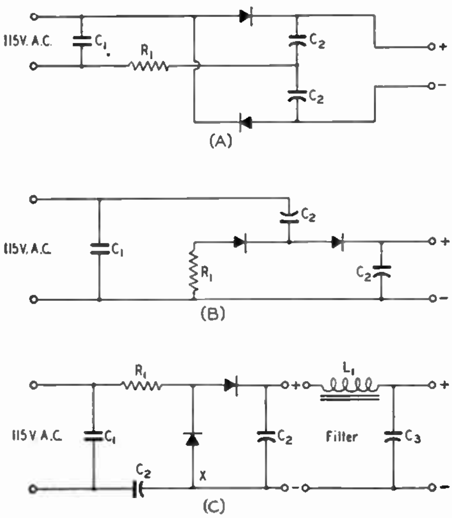


Fig. 7-24 — Voltage-doubling circuits for use with selenium rectifiers.

C_1 — 0.05-μfd. 600-volt paper.
C_2 — 40-μfd 200-volt electrolytic.
C_3 — Filter condenser.
R_1 — 25 to 100 ohms.
L_1 — Filter choke.

shown at B is the most desirable since there is no series condenser. It is a full-wave circuit and there will be very little ripple voltage appearing at the output. On the other hand, the circuit of C has one very desirable feature in that point *X* is common to both condensers in the

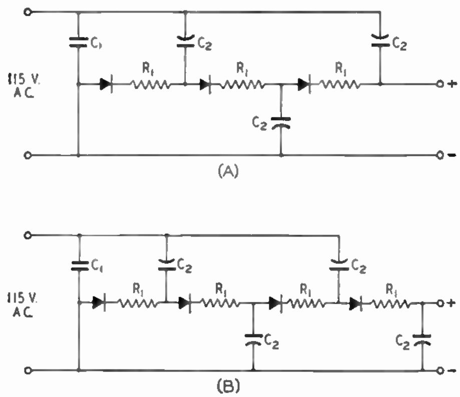


Fig. 7-25 — Selenium-rectifier voltage-tripling and voltage-quadrupling circuits.

C_1 — 0.05-μfd. 600-volt paper.
C_2 — 40-μfd. 450-volt electrolytic.
R_1 — 25 to 100 ohms.

rectifier and also to the first condenser in the filter. This means that a single-unit three-section condenser may be used, saving space. If less than 100 ma. is being used this is the best circuit. The ripple content under these conditions, and the leakage between sections, will not be excessive. These three circuits will find ready application in communications receivers, converters, VFOs, test equipment, etc., and especially in cases where heat has been a problem.

Fig. 7-25A and B shows voltage-tripler and voltage-quadrupler circuits respectively, for use where higher voltages are desired. They are ideal for powering the small rig.

All components are standard. C_1 in all circuits is for "hash" filtering and its value is not critical. A 0.05-μfd. 600-volt-working condenser should serve. All other condensers should be 40-μfd. 200-volt units, except those in the tripler and quadrupler circuits. Those in the circuit of Fig. 7-25 should have a rating of 450 volts working. In the voltage multipliers and in other circuits where a condenser is passing the full current, good condensers should be used because the a.c. ripple mentioned above appears across the condenser and increases as the load increases. If the current is allowed to become too high, it will cause heating and deterioration of the condenser. This can be kept to a minimum by using a capacitor of high value and making sure it is of good make. R_1 should be 25 ohms, but if it is found that the rectifier units are running a little too warm, this value may be increased to as high as 100 ohms, with a corresponding drop in output voltage, of course.

A single-section filter, as shown in Fig. 7-24C, will provide sufficient smoothing for most applications.

Bias Supplies

As discussed in Chapter Six, the chief function of a bias supply for the r.f. stages of a transmitter is that of providing protective bias, although under certain circumstances, a bias supply, or pack, as it is sometimes called, can provide the operating bias if desired.

Simple Bias Packs

Fig. 7-26A shows the diagram of a simple bias supply. R_1 should be the recommended grid leak for the amplifier tube. No grid leak should be used in the transmitter with this type of supply. The output voltage of the supply, when amplifier grid current is not flowing, should be some value between the bias required for plate-current cut-off and the recommended operating bias for the amplifier tube. The transformer peak voltage (1.4 times the r.m.s. value) should not exceed the recommended operating-bias value, otherwise the output voltage of the pack will soar above the operating-bias value when rated grid current flows.

This soaring can be reduced to a considerable extent by the use of a voltage divider across the transformer secondary, as shown at B. Such a system can be used when the transformer voltage is higher than the operating-bias value. The tap on R_2 should be adjusted to give amplifier cut-off bias at the output terminals. The lower the total value of R_2, the less the soaring will be when grid current flows.

A full-wave circuit is shown in Fig. 7-26C. R_3 and R_4 should have the same total resistance and the taps should be adjusted symmetrically. In all cases, the transformer must be designed to furnish the current drawn by these resistors plus the current drawn by R_1.

Regulated Bias Supplies

The inconvenience of the circuits shown in Fig. 7-26 and the difficulty of predicting values in practical application can be avoided in most cases by the use of gaseous voltage-regulator tubes across the output of the bias supply, as shown in Fig. 7-29A. A VR tube with a voltage rating anywhere between the biasing-voltage value which will reduce the input to the amplifier to a safe level when excitation is removed, and the operating value of bias, should be chosen. R_1 is adjusted, without amplifier excitation, until the VR tube ignites and draws about 5 ma. Additional voltage to bring the bias up to the operating value when excitation is applied can be obtained from a grid leak, as discussed in Chapter Six.

Each VR tube will handle 40 ma. of grid current. If the grid current exceeds this value under any condition, similar VR tubes should

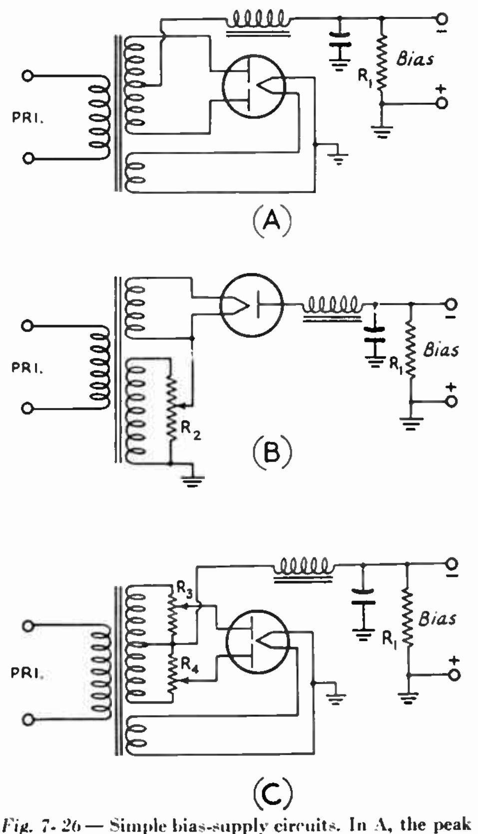


Fig. 7-26 — Simple bias-supply circuits. In A, the peak transformer voltage must not exceed the operating value of bias. The circuits of B (half-wave) and C (full-wave) may be used to reduce transformer voltage to the rectifier. R_1 is the recommended grid-leak resistance.

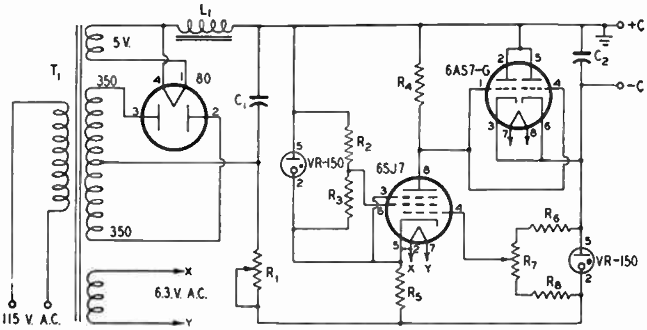


Fig. 7-27 — Circuit diagram of an electronically-regulated bias supply.

C_1 — 20-μfd. 450-volt electrolytic.
C_2 — 20-μfd. 150-volt electrolytic.
R_1 — 5000 ohms, 25 watts.
R_2 — 22,000 ohms, ½ watt.
R_3 — 68,000 ohms, ½ watt.
R_4 — 0.27 megohm, ½ watt.
R_5 — 3000 ohms, 5 watts.
R — 0.12 megohm, ½ watt.
R_7 — 0.1-megohm potentiometer.
R_8 — 27,000 ohms, ½ watt.
L_1 — 20-hy. 50-ma. filter choke.
T_1 — Power transformer: 350 volts r.m.s. each side of center, 50 ma.; 5 volts, 2 amp.; 6.3 volts, 3 amp.

be added in parallel, as shown in Fig. 7-29B, for each 40 ma., or less, of additional grid current. The resistors R_2 are for the purpose of helping to maintain equal currents through each VR tube.

If the voltage rating of a single VR tube is not sufficiently high for the purpose, other VR tubes may be used in series (or series-parallel if required to satisfy grid-current requirements) as shown in Fig. 7-29C and D.

If a single value of fixed bias will serve for more than one stage, the biasing terminal of each such stage may be connected to a single supply of this type, provided only that the total grid current of all stages so connected does not exceed the current rating of the VR tube or tubes. Alternatively, other separate VR-tube branches may be added in any desired combination to the same supply, as shown in Fig. 7-29E, to suit the needs of each stage.

Providing the VR-tube current rating is not exceeded, a series arrangement may be tapped for lower voltage, as shown at F.

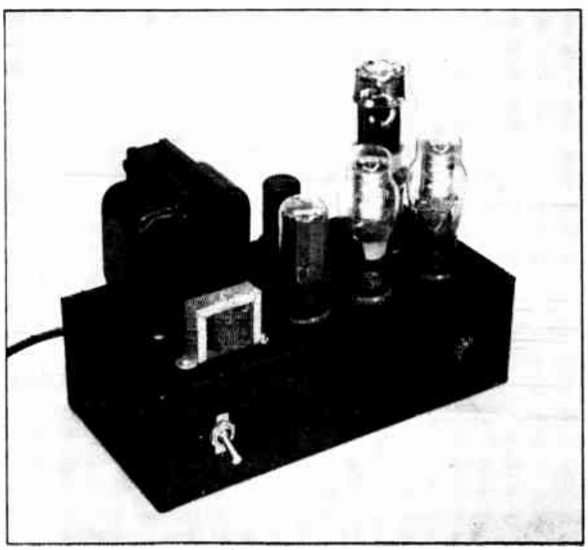
The circuit diagram of an electronically-regulated bias-supply is shown in Fig. 7-27. The output voltage may be adjusted to any value between 20 volts and 80 volts and the unit will handle grid currents up to 200 ma. over the range of 30 to 80 volts, and 100 ma. over the remainder of the range. This will take care of the bias requirements of most tubes used in Class B amplifier service. The regulation will hold to about 0.001 volt per milliampere of grid current. Fig. 7-28 is a photograph of the completed unit.

Other Sources of Biasing Voltage

In some cases, it may be convenient to obtain the biasing voltage from a source other than a separate supply. A half-wave rectifier may be connected with reversed polarization to obtain biasing voltage from a low-voltage plate supply, as shown in Fig. 7-30A. In another arrangement, shown at B, a spare filament winding can be used to operate a filament transformer of similar voltage rating in reverse to obtain a voltage of about 130 from the winding that is customarily the primary. This will be sufficient to operate a VR75 or VR90.

A bias supply of any of the types discussed requires relatively little filtering, if the output-terminal peak voltage does not approach the operating-bias value, because the effect of the supply is entirely or largely "washed out" when grid current flows.

Fig. 7-28 — An electronically-regulated bias supply. Small components are mounted underneath the 3 × 10 × 3-inch chassis. The circuit diagram is shown in Fig. 7-27.



Other Power Considerations

FILAMENT SUPPLY

Except for tubes designed for battery operation, the filaments or heaters of vacuum tubes used in both transmitters and receivers are universally operated on alternating current obtained from the power line through a step-down transformer delivering a secondary voltage equal to the rated voltage of the tubes used. The transformer should be designed to carry the current taken by the number of tubes which may be connected in parallel across it. The filament or heater transformer generally is center-tapped, to provide a balanced circuit for eliminating hum.

For medium- and high-power r.f. stages of transmitters, and for high-power audio stages, it is desirable to use a separate filament transformer for each section of the transmitter, installed near the tube sockets. This avoids the necessity for abnormally large wires to carry the total filament current for all stages without appreciable voltage drop. Maintenance of rated filament voltage is highly important, especially with thoriated-filament tubes, since under- or over-voltage may reduce filament life.

LINE-VOLTAGE ADJUSTMENT

In certain communities trouble is sometimes experienced from fluctuations in line voltage. Usually these fluctuations are caused by a variation in the load on the line and, since most of the variation comes at certain fixed times of the day or night, such as the times when lights are turned on and off at evening, they may be taken care of by the use of a manually-operated compensating device. A simple arrangement is shown in Fig. 7-31A. A toy transformer is used to boost or buck the line voltage as required. The transformer should have a tapped secondary varying between 6 and 20 volts in steps of 2 or 3 volts and its secondary should be capable of carrying the full load current of the entire transmitter, or that portion of it fed by the toy transformer.

The secondary is connected in series with the line voltage and, if the phasing of the windings is correct, the voltage applied to the primaries of the transmitter transformers can be brought up to the rated 115 volts by setting the toy-transformer tap switch on the right tap. If the phasing of the two windings of the toy transformer happens to be reversed, the voltage will be reduced instead of increased. This connection may be used in cases where the line voltage may be above 115 volts. This method is preferable to using a resistor in the primary of a power transformer since it does not affect the voltage regulation as seriously. The circuit of 7-31B illustrates the use of a variable transformer (Variac) for adjusting line voltage to the desired value.

Another scheme by which the primary volt-

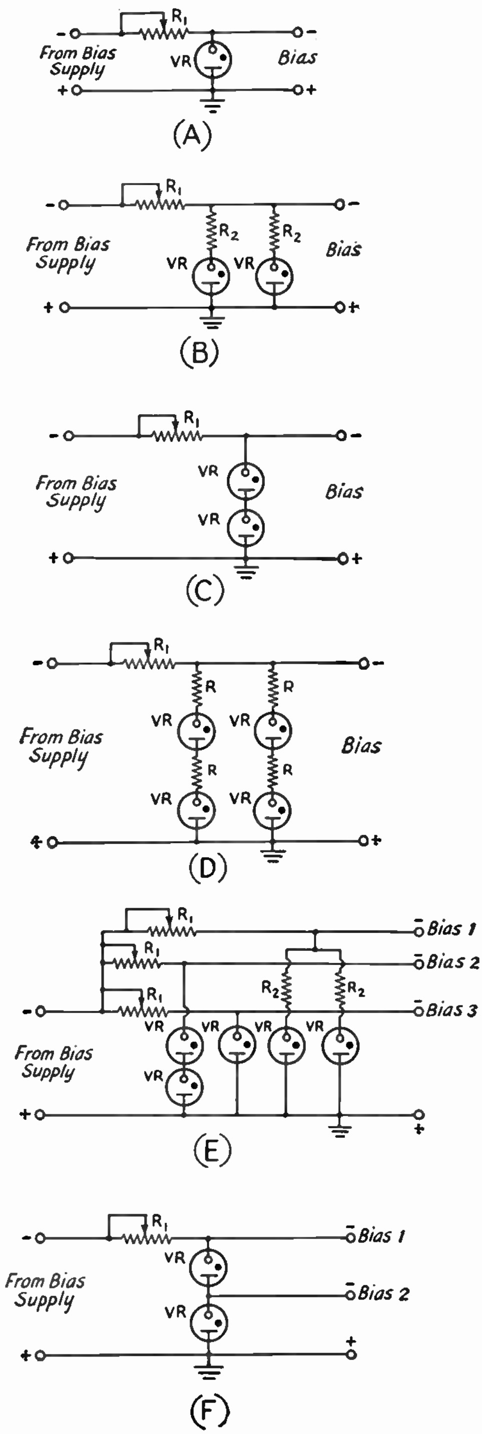


Fig. 7-29 — Illustrating the use of VR tubes in stabilizing protective-bias supplies. R_1 is a resistor whose value is adjusted to limit the current through each VR tube to 5 ma. before amplifier excitation is applied. R and R_2 are current-equalizing resistors of 50 to 100 ohms.

age of each transformer in the transmitter may be adjusted to deliver the desired secondary voltage, with a master control for compensating for changes in line voltage, is described in Fig. 7-32.

This arrangement has the following features:

1) Adjustment of the switch S_1 to make the voltmeter read 105 volts automatically adjusts all transformer primaries to the predetermined correct voltage.

2) The necessity for having all primaries work at the same voltage is eliminated. Thus, 110 volts can be applied to the primary of one transformer, 115 to another, etc.

3) Independent control of the plate transformer is afforded by the tap switch S_2. This permits power-input control and does not require an extra autotransformer.

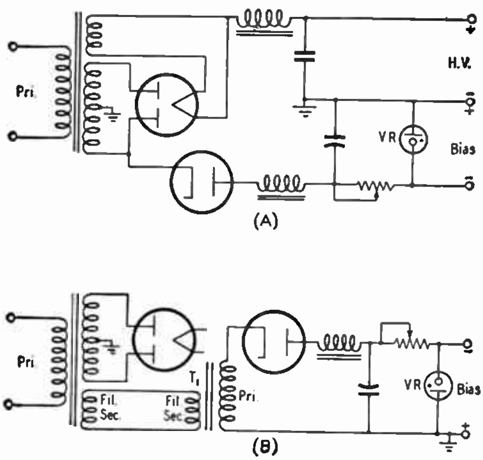


Fig. 7-30 — Convenient means of obtaining biasing voltage. A — From a low-voltage plate supply. B — From spare filament winding. T_1 is a filament transformer, of a voltage output similar to that of the spare filament winding, connected in reverse to give 115 volts r.m.s. output. If cold-cathode or selenium rectifiers are used, no additional filament supply is required.

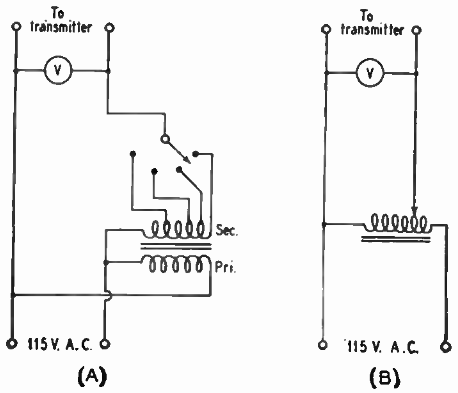


Fig. 7-31 — Two methods of transformer primary control. At A is a tapped toy transformer which may be connected so as to boost or buck the line voltage as required. At B is indicated a variable transformer or autotransformer (Variac) in series with the transformer primaries.

CONSTRUCTION OF POWER SUPPLIES

The length of most leads in a power supply is unimportant, so that the arrangement of components from this consideration is not a factor in construction. More important are the points of good high-voltage insulation, adequate conductor size for filament wiring, proper ventilation for rectifier tubes and — most important of all — safety to the operator. Exposed high-voltage terminals or wiring which might be bumped into accidentally should not be permitted to exist. They should be covered with adequate insulation or placed inaccessible to contact during normal operation and adjustment of the transmitter.

Rectifier filament leads should be kept short to assure proper voltage at the rectifier socket, and the sockets should have good insulation and adequate contact surface. Plate leads to mercury-vapor tubes should be kept short to minimize the radiation of noise.

Where high-voltage wiring must pass through a metal chassis, grommet-lined clearance holes will serve for voltages up to 500 or 750, but ceramic feed-through insulators should be used for higher voltages. Bleeder and voltage-dropping resistors should be placed where they are open to air circulation. Placing them in confined space reduces the power rating.

It is highly preferable from the standpoint of operating convenience to have separate filament transformers for the rectifier tubes, rather than to use combination transformers, such as those used in receivers. This permits the transmitter plate voltage to be switched on without the necessity for waiting for rectifier filaments to come up to temperature after

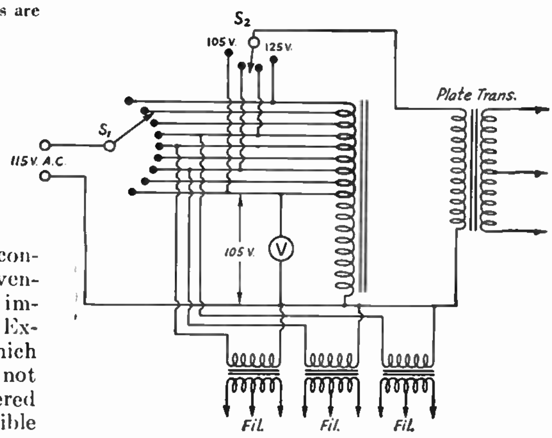


Fig. 7-32 — With this circuit, a single adjustment of the tap switch S_1 places the correct primary voltage on all transformers in the transmitter. Information on constructing a suitable autotransformer at negligible cost is contained in the text. The light winding represents the regular primary winding of a revamped transformer, the heavy winding the voltage-adjusting section.

each time the high voltage has been turned off.

A bleeder resistor with a power rating giving a considerable margin of safety should be used across the output of all transmitter power supplies so that the filter condensers will be discharged when the high-voltage transformer is turned off. To guard against the possibility of danger to the operator should the bleeder resistor burn out without his knowledge, a relay with its winding connected in parallel with the high-voltage transformer primary and its contacts in series with a 1000-ohm resistor across the output of the power supply sometimes is used. The protective relay should be arranged so that the contacts open when the relay is energized.

Emergency and Independent Power Sources

Emergency power supply which operates independently of a.c. lines is available, or can be built in a number of different forms, depending upon the requirements of the service for which it is intended.

The most practical supply for the average individual amateur is one that operates from a 6-volt car storage battery. Such a supply may take the form of a small motor generator (often called a genemotor), a rotary converter, or a vibrator-transformer-rectifier combination.

Dynamotors

A dynamotor differs from a motor generator in that it is a single unit having a double armature winding. One winding serves for the driving motor, while the output voltage is taken from the other. Dynamotors usually are operated from 6-, 12-, 28- or 32-volt storage batteries and deliver from 300 to 1000 volts or more at various current ratings.

Genemotor is a term popularly used when making reference to a dynamotor designed especially for automobile-receiver, sound-truck and similar applications. It has good regulation and efficiency, combined with economy of operation. Standard models of genemotors have ratings ranging from 135 volts at 30 ma. to 300 volts at 200 ma. or 600 volts at 300 ma. The normal efficiency averages around 50 per cent, increasing to better than 60 per cent in the higher-power units. The voltage regulation of a genemotor is comparable to that of well-designed a.c. supplies.

Successful operation of dynamotors and genemotors requires heavy direct leads, mechanical isolation to reduce vibration, and thorough r.f. and ripple filtration. The shafts and bearings should be thoroughly "run in" before regular operation is attempted, and thereafter the tension of the bearings should be checked occasionally to make certain that no looseness has developed.

In mounting the genemotor, the support should be in the form of rubber mounting blocks, or equivalent, to prevent the transmission of vibration mechanically. The frame of the genemotor should be grounded through a heavy flexible connector. The brushes on the high-voltage end of the shaft should be bypassed with 0.002-μfd. mica condensers to a common point on the genemotor frame, preferably to a point inside the end cover close to the brush holders. Short leads are essential. It may prove desirable to shield the entire unit, or even to remove the unit to a distance of three or four feet from the receiver and antenna lead.

When the genemotor is used for receiving, a filter should be used similar to that described for vibrator supplies. A 0.01-μfd. 600-volt (d.c.) paper condenser should be connected in shunt across the output of the genemotor, followed by a 2.5-mh. r.f. choke in the positive high-voltage lead. From this point the output should be run to the receiver power terminals through a smoothing filter using 4- to 8-μfd. condensers and a 15- or 30-henry choke having low d.c. resistance.

A.C.-D.C. Converters

In some instances it is desirable to utilize existing equipment built for 115-volt a.c. operation. To operate such equipment with any of the power sources outlined above would require a considerable amount of rebuilding. This can be obviated by using a rotary converter capable of changing the d.c. from 6-, 12- or 32-volt batteries to 115-volt 60-cycle a.c. Such converter units are built to deliver outputs ranging from 40 to 300 watts, depending upon the battery power available

The conversion efficiency of these units averages about 50 per cent. In appearance and operation they are similar to genemotors of equivalent rating. The over-all efficiency of the converter will be lower, however, because of losses in the a.c. rectifier-filter circuits and the necessity for converting heater (which is supplied directly from the battery in the case of the genemotor) as well as plate power.

Vibrator Power Supplies

The vibrator type of power supply consists of a special step-up transformer combined with a vibrating interrupter (*vibrator*). When the unit is connected to a storage battery, plate power is obtained by passing current from the battery through the primary of the transformer. The circuit is made and reversed rapidly by the vibrator contacts, interrupting the current at regular intervals to give a changing magnetic field which induces a voltage in the secondary. The resulting square-wave d.c. pulses in the primary of the transformer cause an alternating voltage to be developed in the secondary. This high-voltage a.c. in turn is rectified, either by a vacuum-tube rectifier or by an additional synchronized pair of vibrator contacts. The rectified output is pulsating d.c., which may be filtered by ordinary means. The smoothing filter can be a single-section affair, but the filter output capacitance should be fairly large — 16 to 32 μfd.

Fig. 7-33 shows the two types of circuits. At A is shown the *nonsynchronous* type of vibrator. When the battery is disconnected the reed is midway between the two contacts, touching neither. On closing the battery circuit the magnet coil pulls the reed into contact with one contact point, causing current to flow through the lower half of the transformer primary winding. Simultaneously, the magnet coil is short-circuited, deënergizing it, and the reed swings back. Inertia carries the reed into contact with the upper point, causing current to flow through the upper half of the transformer primary. The magnet coil again is energized, and the cycle repeats itself.

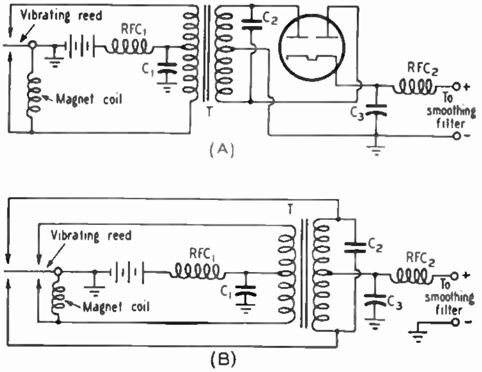


Fig. 7-33 — Basic types of vibrator power-supply circuits. A—Nonsynchronous. B—Synchronous.

The synchronous circuit of Fig. 7-33B is provided with an extra pair of contacts which rectify the secondary output of the transformer, thus eliminating the need for a separate rectifier tube. The secondary center-tap furnishes the positive output terminal when the relative polarities of primary and secondary windings are correct. The proper connections may be determined by experiment.

The buffer condenser, C_2, across the transformer secondary, absorbs the surges that occur on breaking the current, when the magnetic field collapses practically instantaneously and hence causes very high voltages to be induced in the secondary. Without this condenser excessive sparking occurs at the vibrator contacts, shortening the vibrator life. Correct values usually lie between 0.005 and 0.03 μfd., and for 250–300-volt supplies the condenser should be rated at 1500 to 2000 volts d.c. The exact capacitance is critical, and should be determined experimentally. The optimum value is that which results in least battery current for a given rectified d.c. output from the supply. In practice the value can be determined by observing the degree of vibrator sparking as the capacitance is changed. When the system is operating properly there should be practically no sparking at the vibrator contacts. A 5000-ohm resistor in series with C_2 will limit the secondary current to a safe value should the condenser fail.

A more exact check on the operation can be secured with an oscilloscope having a linear sweep circuit that can be synchronized with the vibrator. The vertical plates should be connected across the outside ends of the transformer primary winding to show the input voltage waveshape. Fig. 7-34C shows an idealized trace of the optimum waveform when the buffer capacitor is adjusted to give proper operation throughout the life of the vibrator. The horizontal lines in the trace represent the voltage during the time the vibrator contacts are closed, which should be approximately 90 per cent of the total time. When the contacts are open the trace should be partly tilted and partly vertical, the tilted part being 60 per cent of the total connecting trace. The oscilloscope will show readily the effect of the buffer condenser on the percentage of tilt. In actual patterns the horizontal sections are likely to droop somewhat because of the resistance drop in the battery leads as the current builds up through the primary inductance (Fig. 7-34D). Trace E shows the result of insufficient buffering capacitance, while too much buffering capacitance will show a slow build-up in voltage with rounded corners evident in the trace. Figs.

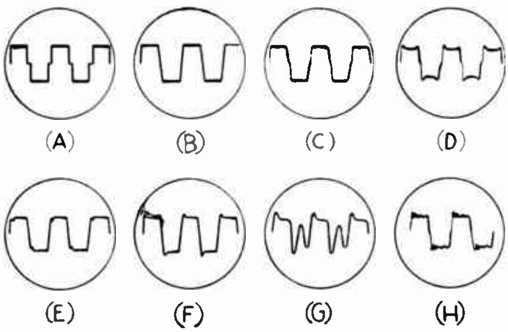


Fig. 7-34 — Characteristic vibrator waveforms as viewed on the oscilloscope. A, ideal theoretical trace for resistive load; current flow stops instantly when vibrator contacts open and resumes approximately 1 microsecond later (for standard 115-cycle vibration frequency) after interrupter arm moves across for the next half-cycle. B, ideal practical waveform for inductive load (transformer primary) with correct buffer capacitance. C, practical approximation of B for loaded nonsynchronous vibrator. D, satisfactory practical trace for synchronous (self-rectifying) vibrator under load; the peaks result from voltage drop in the primary when the secondary load is connected, not from faulty operation.

Faulty operation is indicated in E through H: E, effect of insufficient buffering capacitance (not to be mistaken for "bouncing" of contacts). The opposite condition — excessive buffering capacitance — is indicated by slow build-up with rounded corners, especially on "open." F, overclosure caused by too-small buffer condenser (same condition as in E) with vibrator unloaded. G, "skipping" of worn-out or misadjusted vibrator, with interrupter making poor contact on one side. H, "bouncing" resulting from worn-out contacts or sluggish reed. G and H usually call for replacement of the vibrator.

7-34G and H indicate a worn or improperly-adjusted vibrator.

"Hash" Elimination

Sparking at the vibrator contacts causes r.f. interference ("hash," which can be distinguished from hum by its harsh, sharper pitch) when used with a receiver. To minimize this, r.f. filters are incorporated, consisting of RFC_1 and C_1 in the battery circuit, and RFC_2 with C_3 in the d.c. output circuit.

Equally as important as the hash filter is thorough shielding of the power supply and its connecting leads, since even a small piece of wire or metal will radiate enough r.f. to cause interference in a sensitive receiver.

Testing in connection with hash elimination should be carried out with the supply operating a receiver. Since the interference usually is picked up on the receiving-antenna leads by radiation from the supply itself and from the battery leads, it is advisable to keep the supply and battery as far from the receiver as the connecting cables will permit. Three or four feet should be ample. The microphone cord likewise should be kept away from the supply and leads.

The power supply should be built on a metal chassis, with all unshielded parts underneath. A bottom plate to complete the shielding is advisable. The transformer case, vibrator cover and the metal shell of the tube all should be grounded to the chassis. If a glass tube is used it should be enclosed in a tube shield. The battery leads should be evenly twisted, since these leads are more likely to radiate hash than any other part of a well-shielded supply. Experimenting with different values in the hash filters should come *after* radiation from the battery leads has been reduced to a minimum. Shielding the leads is not particularly helpful.

PRACTICAL VIBRATOR-SUPPLY CIRCUITS

A vibrator-type power supply may be designed to operate from a six-volt storage battery only, or in a combination unit which may be operated interchangeably from either battery or 115 volts a.c.

Typical circuits are shown in Fig. 7-35. The one shown at A is the simplest, although it operates from a 6-volt d.c. source only. S_1 turns the high voltage on and off.

The circuit of B provides for either 6-volt d.c. or 115-volt a.c. operation with a dual-primary transformer. S_2 is the a.c. on-off switch while S_3 switches the heater of the 6X5 rectifier from the storage battery to the 6.3-volt winding on the transformer. Filament supply for the transmitter or receiver is switched by shifting the power plug to the correct output socket, X when operating from a 6-volt d.c. source, and Y when 115-volt a.c. input is used.

The circuit of Fig. 7-35C may be used when a dual-primary transformer is not available. The filter is switched from one rectifier

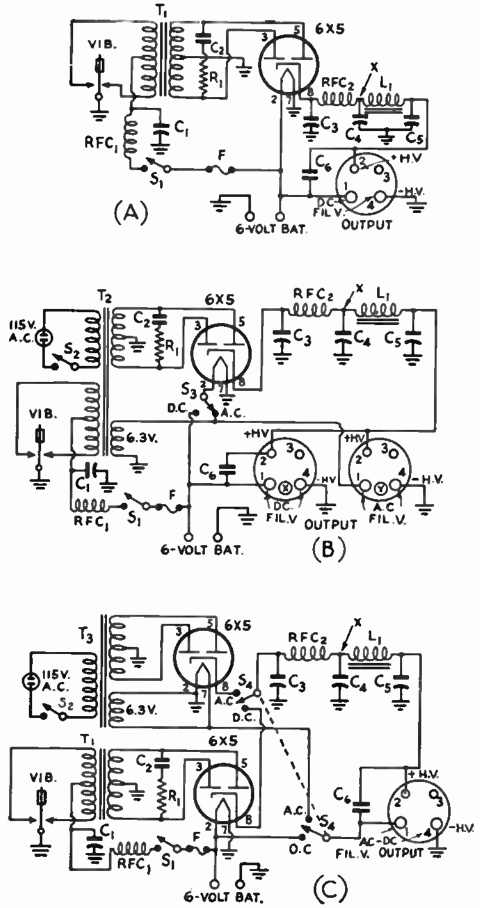


Fig. 7-35 — Typical vibrator-transformer power-supply circuits. The circuit at A shows a simple arrangement for 6-volt d.c. input; the one at B illustrates the use of a combination transformer for operation from either 6 volts d.c. or 115 volts, a.c. The circuit of C is similar to that of B but uses separate transformers.

C_1 — 0.5-µfd. paper, 50-volt rating or higher.
C_2 — 0.005 to 0.01 µfd., 1600 volts.
C_3 — 0.01-µfd. 600-volt paper.
C_4 — 8-µfd. 450-volt electrolytic.
C_5 — 32-µfd. 450-volt electrolytic.
C_6 — 100-µµfd. mica.
R_1 — 4700 ohms, ½ or 1 watt.
L_1 — 10/12-henry 100-ma. filter choke, not over 100 ohms (Stancor C-2303 or equivalent).
F — 15-ampere fuse.
RFC_1 — 55 turns No. 12 on 1-inch form, close-wound.
RFC_2 — 2.5-mh. r.f. choke.
S_1 — S.p.s.t. toggle — battery switch.
S_2 — S.p.s.t. toggle — a.c. power switch.
S_3 — S.p.d.t. toggle — rectifier-heater change-over switch.
S_4 — D.p.d.t. toggle — a.c.-d.c. switch.
T_1 — Vibrator transformer.
T_2 — Special vibrator transformer with 115-volt and 6-volt primaries, to give approximately 300 volts at 100 ma. d.c. (Stancor P-6166 or equivalent).
T_3 — 275 to 300 volts each side of center-tap, 100 to 150 ma.; 6.3-volt filament.
VIB — Vibrator unit (Mallory 500P, 294, etc.)
X — Insert a series resistor of suitable value to drop the output voltage to 300 at 100 ma. load, if necessary. If transformer gives over 300 volts d.c., a second filter choke may be used to give additional voltage drop as well as more smoothing.

output to the other by means of the d.p.d.t. switch, S_4, which also shifts filament connections from a.c. to d.c. The filter section of the switch could be eliminated if desired by connecting the filtering circuit permanently to the output terminals of both rectifiers and removing the unused rectifier tube from its socket. Similarly, the filament section of S_4 could be dispensed with by providing two output sockets as in the circuit at B. If a separate rectifier filament winding is available on T_3, directly-heated rectifier types may be substituted for the 6X5 in the a.c. supply. In some cases where the required filament windings are not available, a rectifier of the cold-cathode type, such as the 0Z4, which requires no heater voltage, sometimes may be used to advantage.

If suitable filament windings are available, a regular a.c. transformer will make an acceptable substitute for a vibrator transformer. If the a.c. transformer has two 6.3-volt windings, they may be connected in series, their junction forming the required center-tap. A 6.3-volt and a 5-volt winding may be used in a similar manner even though the junction of the two windings does not provide an accurate center-tap. A better center-tap may be obtained if a 2.5-volt winding also is available, since half of this winding may be connected in series with the 5-volt winding to give 6.25 volts.

R.f. filters for reducing hash are incorporated in both primary and secondary circuits. The secondary filter consists of a 0.01-μfd. paper condenser directly across the rectifier output, with a 2.5-mh. r.f. choke in series ahead of the smoothing filter. In the primary circuit a low-inductance choke and high-capacitance condenser are needed because of the low impedance of the circuit. A choke of the specifications given should be adequate, but if there is trouble with hash it may be beneficial to experiment with other sizes. The wire should be large — No. 12, preferably, or No. 14 as a minimum. Manufactured chokes such as the Mallory RF583 are more compact and give higher inductance for a given resistance because they are bank-wound, and may be substituted if obtainable. C_1 should be at least 0.5 μfd.; even more capacitance may help in bad cases of hash.

The smoothing filter for battery operation can be a single-section affair, but there will be some hum (readily distinguishable from hash because of its deeper pitch) unless the filter output capacitance is fairly large — 16 to 32 μfd.

The compactness of selenium rectifiers and the fact that they do not require filament voltage make them particularly suited to compact lightweight power supplies for portable-emergency work.

Fig. 7-36 shows the circuit of a vibrator pack that will deliver an output voltage of 400 at 200 ma. It will work with either 115-volt ac. or 6-volt battery input. The circuit is that of the familiar voltage tripler whose d.c. output voltage is, as a rough approximation, three times the peak voltage delivered by the transformer or line. An interesting feature of the circuit is the fact that the single transformer serves as the vibrator transformer when operating from 6-volt d.c. supply and as the filament transformer when operating from an a.c. line. This is accomplished without complicated switching.

The vibrator transformer, T_1, is a dual-secondary 6.3-volt filament transformer connected in reverse. It may also consist of two single transformers of the same type with their primaries connected in series and secondaries in parallel, both windings being properly polarized. In either event, the filament windings must have a rating of 10 amperes if the full load current of 200 ma. is to be used. Some excellent surplus transformers that will handle the required current are now available on the surplus market. The vibrator also must be capable of handling the current. The hash-filter choke, L_1, must carry a current of 20 amperes.

The following table shows the output voltage to be expected at various load currents, depending upon the size of condensers used at C_1, C_2 and C_3.

C_1, C_2, C_3 (μfd.)	*Output Voltage at* 50 ma.	100 ma.	150 ma.	200 ma.
60	455	430	415	395
40	425	390	360	330
20	400	340	285	225

In operating the supply from an a.c. line, it is always wise to determine the plug polarity with respect to ground. Otherwise the rectifier part of the circuit and the transformer circuit can-

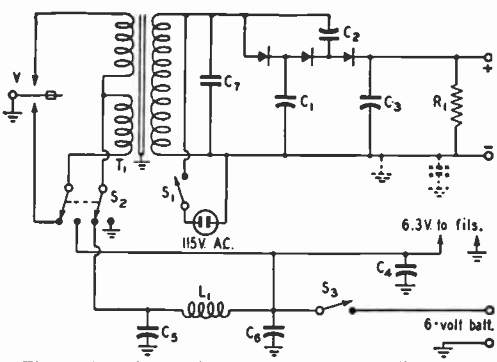


Fig. 7-36 — Circuit diagram of a compact vibrator-a.c. portable power supply suggested by W9CO.

C_1 — 60-μfd. 200-volt electrolytic.
C_2 — 60-μfd. 400-volt electrolytic.
C_3 — 60-μfd. 600-volt electrolytic.
C_4 — 25-μfd. 25-volt electrolytic.
C_5, C_6 — 0.5-μfd. 25-volt paper.
C_7 — 0.007-μfd. 1500-volt paper.
R_1 — 25,000 ohms, 10 watts.
L_1 — 25-μhy. 20-amp. choke.
S_1 — 115-volt toggle switch.
S_2 — D.p.d.t. heavy-duty knife switch.
S_3 — 25-amp. s.p.s.t. switch.
T_1 — See text.
V — Heavy-duty vibrator.

TABLE 7-II PLATE-BATTERY SERVICE HOURS

Estimated to 34-volt end-point per nominal 45-volt section.
Based on intermittent use of 3 to 4 hours daily at room temp. of 70° F.
(For batteries manufactured in U. S. A. only.)

Manufacturer's Type No.		Weight		Current Drain in Ma.											
Burgess	Eveready	Lb.	Oz.	5	10	15	20	25	30	40	50	60	75	100	150
—	758	14	8	Suggested current range = 7 to 12 Ma.											
21308	—	12	8	1600	1100	690	490	—	300	200	—	130	—	60	30
10308	—	11	4	1300	750	520	350	—	—	130	—	90	—	45	22
—	754	6	8	Suggested current range = 5 to 15 Ma.											
2308	—	8	3	1100	500	330	200	—	150	80	—	43	—	—	—
—	487	4	2	Suggested current range = 7 to 12 Ma.											
B30	—	2	8	350	170	90	50	—	21	17	—	—	—	—	—
A30	—	2	—	260	100	48	28	—	17	7	—	—	—	—	—
—	482	1	14	400	200	122	80	—	—	—	—	—	—	—	—
Z30N	—	1	4	155	70	30	20	15	9.5	—	—	—	—	—	—
—	467	—	12	82	30	—	—	—	—	—	—	—	—	—	—
—	738	1	2	160	70	30	20	10	7	—	—	—	—	—	—
W30FL	—	—	11	70	20	12	7	—	3.5	—	—	—	—	—	—
—	455	—	8	82	30	—	—	—	—	—	—	—	—	—	—
XX30	—	—	9	70	20	12	7	—	3.5	—	—	—	—	—	—

not be connected to actual ground except through by-pass condensers.

Vibrator-Supply Construction

A typical example of vibrator-supply construction is shown in the photograph of Fig. 7-37.

This model makes use of separate power transformers for 115-volt a.c. and 6-volt d.c. operation, the single rectifier tube being shifted from one octal socket to the other when the change from a.c. to d.c. operation is made. The components are assembled on a 5 × 10 × 3-inch steel chassis. The two transformers are flush-mounting type requiring cut-outs in the chassis. Three socket holes are required — one for the 4-prong socket for the vibrator and two octal sockets for the rectifier. The a.c. line cord and battery and power-output leads are brought out at the rear.

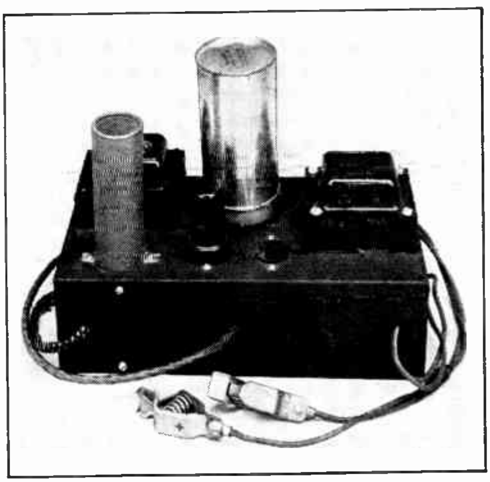


Fig. 7-37 — A typical combination a.c.-d.c. power pack for low-power emergency work. The two transformers are flush-mounted at either end of the chassis. The filter condenser is at the left, the two rectifier sockets at the center and the vibrator to the rear.

GASOLINE-ENGINE DRIVEN GENERATORS

For higher-power installations, such as for communications control centers during emergencies, the most practical form of independent power supply is the gasoline-engine driven generator which provides standard 115-volt 60-cycle supply.

Such generators are ordinarily rated at a minimum of 250 or 300 watts. They are available up to two kilowatts, or big enough to handle the highest-power amateur rig. Most are arranged to charge automatically an auxiliary 6- or 12-volt battery used in starting. Fitted with self-starters and adequate mufflers and filters, they represent a high order of performance and efficiency. Many of the larger models are liquid-cooled, and they will operate continuously at full load.

A variant on the generator idea is the use of fan-belt drive. The disadvantage of requiring that the automobile must be running throughout the operating period has not led to general popularity of this idea among amateurs. Such generators are similar in construction and capacity to the small gas-driven units.

The output frequency of an engine-driven generator must fall between the relatively narrow limits of 50 to 60 cycles if standard 60-cycle transformers are to operate efficiently from this source. A 60-cycle electric clock provides a means of checking the output frequency with a fair degree of accuracy. The clock is connected across the output of the generator and the second hand is checked closely against the second hand of a watch. The speed of the engine is adjusted until the two second hands are in synchronism. If a 50-cycle clock is used to check a 60-cycle generator, it should be remembered that one revolution of the second hand will be made in 50 seconds and the clock will gain 4.8 hours in each 24 hours.

Output voltage should be checked with a

TABLE 7-III — FILAMENT-BATTERY SERVICE HOURS

Estimated to 1-volt end-point per nominal 1.5-volt unit. Based on intermittent use of 3 to 4 hours per day at room temperature. (For batteries manufactured in U. S. A. only.)

Manufacturer's Type No.		Weight		Voltage	Current Drain in Ma.											
Burgess	Eveready	Lb.	Oz.		30	50	60	120	150	175	100	200	240	250	300	350
—	A-1300	8	4	1.25	—	—	—	—	2000	1715	1500	1333	1250	1200	1000	054
—	740	6	4	1.5	—	—	—	—	—	—	—	870	—	—	—	—
—	741[1]	2	13	1.5	—	—	—	—	—	—	—	460	—	—	270	—
—	743	2	1	1.5	—	—	—	—	—	—	—	300	—	225	175	—
—	742	1	6	1.5	—	—	—	—	—	—	—	170	—	120	90	—
8F[2]	—	2	10	1.5	—	—	1100	600	450	—	—	400	—	320	230	190
4F	—	1	4	1.5	—	—	600	340	230	—	—	160	—	110	95	60
—	A-2300	11	—	2.5	—	—	—	—	2000	1715	1500	1333	1250	1200	1000	054
20F2	—	13	12	3.0	—	—	—	—	1100	—	—	850	—	775	600	500
2F2H	—	1	6	3.0	600	—	340	130	95	—	—	60	—	42	30	—
2F2BP[3]	—	1	5	3.0	600	—	340	130	95	—	—	60	—	42	30	—
F2DP	—	—	12	3.0	340	—	130	45	30	—	—	—	—	—	—	—
G3[4]	—	1	5	4.5	370	200	150	50	35	—	—	—	—	—	—	—
—	746	1	4	4.5	—	225	—	—	—	—	—	—	—	—	—	—
—	718[5]	2	13	6.0	—	415	—	—	—	—	—	—	—	—	—	—
F4PI	—	1	6	6.0	340	150	130	45	30	—	—	—	—	—	—	—

[1] Same life figures apply to 745, wt. 2 lb. 13 oz.
[2] Same life figures apply to 8FL, wt. 2 lb. 15 oz.
[3] Same life figures apply to 2F4, volts 6, wt. 2 lb. 11 oz.
[4] Same life figures apply to G5, volts 7½, wt. 2 lb. 2 oz.
[5] Same life figures apply to 747, wt. 2 lb. 13 oz.

If batteries of another make are to be used, locate ones of similar size and weight on these tables and comparable performance may be expected.

voltmeter since a standard 115-volt lamp bulb, which is sometimes used for this purpose, is very inaccurate. Tests have shown that what appears to be normal brilliance in the lamp may occur at voltages as high as 150 if the check is made in bright sunlight.

Noise Elimination

Electrical noise which may interfere with receivers operating from engine-driven a.c. generators may be reduced or eliminated by taking proper precautions. The most important point is that of grounding the frame of the generator *and* one side of the output. The ground lead should be short to be effective, otherwise grounding may actually increase the noise. A water pipe may be used if a short connection can be made near the point where the pipe enters the ground, otherwise a good separate ground should be provided.

The next step is to loosen the brush-holder

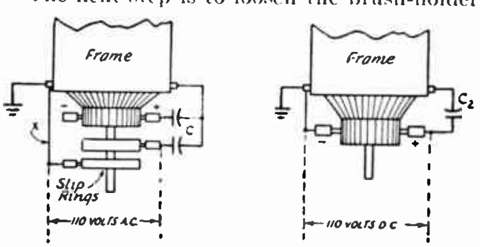


Fig. 7-38 — Connections used for eliminating interference from gas-driven generator plants. C should be 1 μfd., 300 volts, paper, while C_2 may be 1 μfd. with a voltage rating of twice the d.c. output voltage delivered by the generator. X indicates an added connection between the slip ring on the grounded side of the line and the generator frame.

locks and slowly shift the position of the brushes while checking for noise with the receiver. Usually a point will be found (almost always different from the factory setting) where there is a marked decrease in noise.

From this point on, if necessary, by-pass condensers from various brush holders to the frame, as shown in Fig. 7-38, will bring the hash down to within 10 to 15 per cent of its original intensity, if not entirely eliminating it. Most of the remaining noise will be reduced still further if the high-power audio stages are cut out and a pair of headphones is connected into the second detector.

POWER FOR PORTABLES

Dry-cell batteries are the only practical source of supply for equipment which must be transported on foot. From certain considerations they may also be the best source of voltage for a receiver whose filaments may be operated from a storage battery, since no problem of noise filtering is involved.

Their disadvantages are weight, high cost, and limited current capability. In addition, they will lose their power even when not in use, if allowed to stand idle for periods of a year or more. This makes them uneconomical if not used more or less continuously.

Tables 7-II and 7-III give service life of representative types of batteries for various current drains, based on intermittent service simulating typical operation. The continuous-service life will be somewhat greater at very low current drains and from one-half to two-thirds the intermittent life at higher drains.

Keying and Break-In

If the proper keying of a transmitter entailed only the ability to turn on and off the output, keying would be a simple matter. Unfortunately, perfect keying is as difficult to obtain as perfect voice quality, and so is not a matter to be dismissed lightly. The keying of a transmitter can be considered satisfactory if the power output is reduced to zero with the key open, or "up," and reaches full output when the key is closed, or "down." The keying system should accomplish this without producing objectionable transients or "clicks," which cause interference with other amateur stations and with local broadcast reception. Furthermore, the keying process should cause no "chirp," which means that the transmitter output frequency should not be affected by the keying process.

Keying Principles and Characteristics

Back-Wave

When the transmitter output is not reduced to zero under key-up conditions, the signal is said to have a **back-wave.** If the amount of back-wave is appreciable, the keying will be difficult to read. A pronounced back-wave may result when the amplifier feeding the antenna is keyed, as a result of the excitation energy feeding through an incompletely-neutralized stage. Magnetic coupling between antenna coils and one of the driver stages on the operating frequency is also a cause of back-wave. Direct radiation from a driver stage ahead of the keyed stage will result in a back-wave, but this type is generally heard only within a few miles of the transmitter, unless the driver stage is fairly high-powered.

A back-wave also may be radiated if the keying system does not reduce the input to the keyed stage to zero during keying spaces. This trouble will not occur in keying systems that completely cut off the plate voltage when the key is open. It will occur in grid-block keying systems if the blocking voltage is not great enough, or in power-supply primary keying systems if only the final-stage power-supply primary is keyed. A vacuum-tube keyer will give a back-wave if the "open" key resistance is too low.

Key Clicks

If a transmitter is keyed in such a manner that the power output rises instantly to its full value or drops immediately to zero, the resultant short rise and decay times produce signals (at the times of closing and opening the key) extending from the signal frequency to several hundred kilocycles on either side. These signals are called "key clicks," and they will cause interference to other amateurs and other services. Consequently, keying systems must be used that increase the rise and decay times of the keyed characters, since this results in less click energy removed from the signal frequency.

The simple process of making and breaking any circuit with current flowing through it will produce a brief burst of r.f. energy. This effect can be noticed in a radio receiver when an electric light or other appliance in the house is turned on or off. It is, therefore, not only necessary to delay the rise and decay times of the keyed transmitter to prevent interference with other services, but it may be necessary to filter the r.f. energy generated at the key contacts if this energy is found to interfere with broadcast reception in the amateur's house or vicinity. This interference is also called "key clicks."

Getting back to the discussion of rise and decay times, tests have shown that practically all operators prefer to copy a signal that is "solid" on the "make" end of each dot or dash; i.e., one that does not build up too slowly but just slowly enough to have a slight click when the key is closed. On the other hand, the most-pleasing and least-difficult signal to copy, particularly at high speeds, is one that has a fairly soft "break" characteristic; i.e., one that has practically no click as the key is opened. A signal with heavy clicks on both make and break is difficult to copy at high speeds and also causes considerable interference. If it is too "soft" the dots and dashes will tend to run together and the characters will be difficult to copy. The keying should be adjusted so that for all normal hand speeds (15 to 35 w.p.m.) the readability will be satisfactory without causing unnecessary interference to the reception of other signals near the transmitter frequency.

Chirps

Keying should have no effect upon the frequency of the transmitter. In many cases where sufficient pains have not been taken, keying will cause a frequency change, or "chirp," at the instant of opening or closing the key. The resultant signal is unpleasant and, in cases of extreme chirp, difficult to copy. Multistage transmitters keyed in a stage following the oscillator are generally free from chirp, unless the keying causes line-voltage changes which in turn affect the oscillator frequency. When the oscillator is keyed, as is done for "break-in" operation, particular care must be taken to insure that the signal does not have keying chirps.

Break-In Operation

In code transmission, there are intervals between dots and dashes, and slightly longer intervals between letters and words, when no power is being radiated by the transmitter. If the receiver can be made to operate at normal sensitivity during these intervals, it is possible for the receiving operator to signal the transmitting operator, by holding his key down. This is useful during the handling of messages, since the receiving operator can immediately signal the transmitting operator if he misses part of the message. It is also useful in reducing the time necessary for calling in answer to a "CQ." The ability to hear signals during the short "key-up" intervals is called **break-in operation.**

Selecting the Stage To Key

It is highly advantageous from an operating standpoint to design the c.w. transmitter for break-in operation. In most cases this requires that the oscillator be keyed, since a continuously-running oscillator will create interference in the receiver and prevent break-in on or near one's own frequency. On the other hand, it is easier to avoid a chirpy signal by keying a stage or two following the oscillator. Since the effect of a chirp is multiplied with frequency, it is quite difficult to obtain chirpless oscillator keying in the 14- and 28-Mc. bands. In any case, however, the stages following the keyed stage (or stages) must be provided with sufficient fixed bias to limit the plate currents to safe values when the key is up and the tubes are receiving no excitation voltage. Complete cut-off reduces the possibility of a back-wave if a stage other than the oscillator is keyed, but the keying waveform is not well preserved and some clicks can be introduced, even though the keyed stage itself produces no clicks. *It is a good general rule to bias the tubes following the keyed stage so that they draw a key-up current of about 5 per cent of the normal key-down value.*

The power broken by the key is an important consideration, both from the standpoint of safety to the operator and that of sparking at the key contacts. Keying of the oscillator or a low-power stage is favorable on both counts. The use of a keying relay is recommended when a high-power circuit is keyed.

Keying Circuits

Only general circuits can be shown for keying, since the final decision on where and how to key rests with the amateur and depends upon the power level and type of operation.

● PLATE-CIRCUIT KEYING

Any stage of the transmitter can be keyed by opening and closing the plate-power circuit. Fig. 8-1 shows how the key can be connected to key the plate circuit (A) or the screen circuit (B). The circuit of Fig. 8-1A shows the key in the negative power lead, although it could be placed in the positive lead, at the point marked "X." Either system is recommended only for low-voltage circuits, of the order of 300 or less, unless a relay is used, because of the danger of accidental electrical shock.

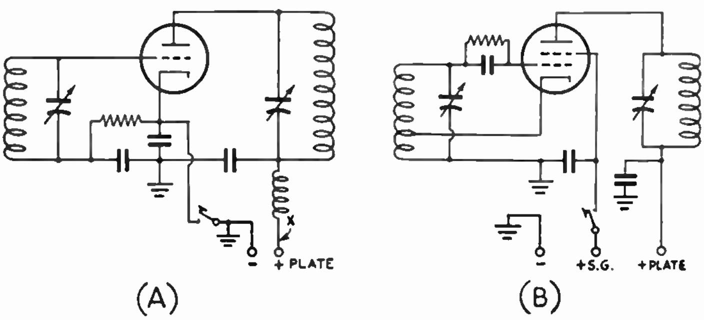


Fig. 8-1 — Plate-circuit keying is shown at A, and screen-grid keying is shown at B. Oscillator circuits are shown in both cases, but the same keying methods can be used with amplifier circuits. Notice the similarity between A and Fig. 8-5 — the only difference is in the way the grid return is connected.

Fig. 8-1B shows the key in the screen lead of an electron-coupled oscillator, and can be considered a variation of 8-1A that has the desirable advantage of breaking less current at a lower voltage.

Both of the circuits shown in Fig. 8-1 respond well to the use of key-click filters, and are particularly suitable for use with crystal- and self-controlled oscillators, which are generally operated at low voltage and low power.

In any transmitter where a driver stage requires the same supply voltage as the screen of the driven stage, the positive lead to the driver stage and to the screen grid of the amplifier can be keyed simultaneously, with excellent results. Usually no fixed bias will be required on the grid of the amplifier, since the key-up plate current will have a low value.

Generally an oscillator will operate at a very low plate voltage, but some refuse to. In the case of the latter, an improvement in keying can sometimes be obtained by using a high value of resistance across the key that will permit the oscillator to draw some plate current (without oscillating). No one value of resistance can be recommended, since every case will be different, but several different values of resistance should be tried, increasing in value until the oscillator stops.

PRIMARY KEYING

A popular method of keying high-powered amplifiers is shown in Fig. 8-2. In its simplest form, as shown in 8-2A, it consists of keying the primary of the plate transformer supplying power to one or more of the driver stages. It has the advantage that the filter, *LC*, acts as a keying filter and prevents clicks. However, too much filter cannot be used or the keying will be too soft, and a single section is all that can normally be used. Since this will introduce some a.c. modulation on the keyed stages, it is essential that the amplifier driven by the keyed stage have sufficient excitation to operate as a Class C amplifier, which tends to eliminate the modulation existing in the excitation voltage. Primary keying of the final plate power supply alone is not recommended, since it is practically impossible to comply with FCC regulations about "adequately-filtered power supply" and still avoid keying that is too soft.

Primary keying of the driver power supply requires that the following amplifier stage (or stages) be biased to prevent excessive current under key-up conditions. If this bias exceeds the cut-off value for the tube (or tubes) a slightly more elaborate version of primary keying can be used, as shown in Fig. 8-2B. The primaries of both driver and final-amplifier plate supplies are keyed, and the system has the advantage that the final-amplifier plate voltage remains substantially constant under key-up or key-down conditions, and thus no clicks can be introduced by the sudden changes in final-amplifier plate voltage as the excitation is applied or removed. The final-

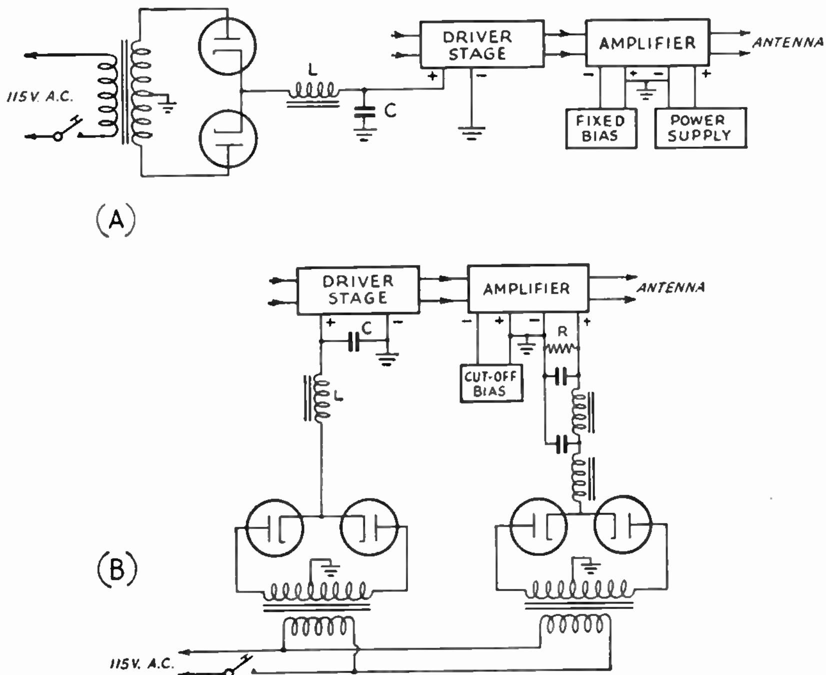


Fig. 8-2 — Primary-keying circuits. The circuit at A shows primary keying of the driver-stage (or stages) power supply, followed by an amplifier biased to or close to cut-off. The circuit of B uses primary keying of both driver and final supplies, and has the advantage that the key-up and key-down voltages on the final amplifier remain substantially constant, thus eliminating the chance of clicks being introduced by the final-amplifier plate-supply regulation.

In either case, *L* and *C* should be as small as possible, consistent with sufficient filtering and rectifier-tube limits. *R* in B need be only about 1000 ohms per volt. If a plate voltmeter is used, the bleed through it is sufficient, since the only function is to remove any long-standing charge from the power supply. A heavy bleed current will reduce the effectiveness of the keying system. See text for other bleeder suggestions.

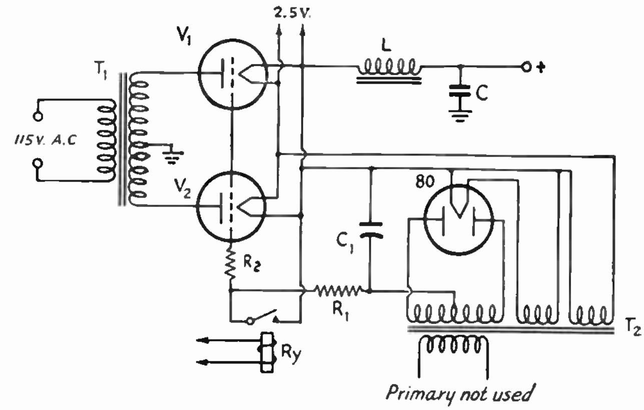


Fig. 8-3 — Grid-controlled rectifier keying. Circuit is similar to Fig. 8-2, and the values of *L* and *C* are the same. A well-insulated keying relay, *Ry*, is used to control the bias on the rectifiers V_1 and V_2. The bias voltage is obtained from a small receiver power-supply transformer T_2, the 80 rectifier, and filter condenser C_1. T_2 does not need to be insulated for the full plate-supply voltage (obtained from T_1) because it is excited from the filament transformer for V_1 and V_2. It should be well insulated to ground, however. R_1 limits the short-circuit on the bias supply and can be approximately 50,000 ohms in value.

amplifier plate supply will remain charged for several minutes after the last transmission, however, and extreme caution must be exercised. As a safety measure, the final-amplifier power supply can be discharged by a relay that shorts the supply through a 1000-ohm resistor, or the bias can be removed and the final-amplifier tube will discharge the power supply.

The keying system shown in Fig. 8-2B has been used to key an entire transmitter for break-in operation. The oscillator and multiplier/driver stages take their plate power from the supply with the small filter, while the final amplifier is powered from the heavily-filtered supply. It is essential, however, in a transmitter keyed for break-in in this manner, that the oscillator be free from chirp, and this point should be checked carefully before using the system on the air.

In using primary keying up to several hundred watts, direct keying in the primary circuit is satisfactory. For higher powers, however, a suitable keying relay should be used, because of the arcing at the contacts.

Fig. 8-3 shows **grid-controlled rectifier tubes** in the power supply. By applying suitable bias to the tubes when the key is up, no current flows through the tubes. When the key is closed, the bias is removed and the tubes conduct. The system can be used in the same way that primary keying was used in Fig. 8-2A and B. This system is used only in high-powered high-voltage supplies.

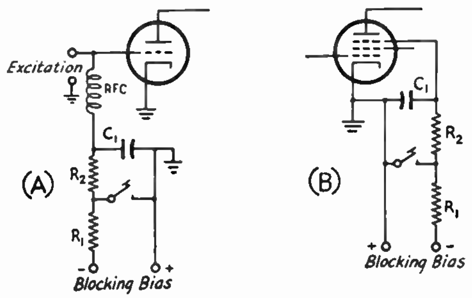


Fig. 8-4 — Blocked-grid keying. R_1, the current-limiting resistor, should have a value of about 50,000 ohms. C_1 may have a capacity of 0.1 to 1 μfd., depending upon the keying characteristic desired. R_2 is the normal value of grid leak for the tube.

● BLOCKED-GRID KEYING

An amplifier tube can be keyed by applying sufficient negative bias voltage to the control or suppressor grid to cut off plate-current flow when the key is up, and by removing this blocking bias when the key is down. When the bias is applied to the control grid, its value will be considerably higher than the nominal cut-off bias for the tube, since the r.f. excitation voltage must be overcome. The fundamental circuits are shown in Fig. 8-4A and B. The circuits can be applied to oscillator tubes as well as amplifiers. Suppressor-grid keying will not completely turn off a Tri-tet crystal oscillator or electron-coupled self-controlled oscillator, and is likely to cause serious chirps with the latter.

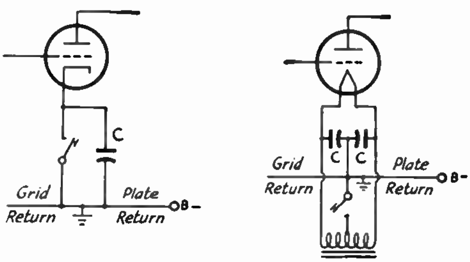


Fig. 8-5 — Cathode and center-tap keying. The condensers *C* are r.f. by-pass condensers. Their capacity is not critical, values of 0.001 to 0.01 μfd. ordinarily being used.

In both circuits the key is connected in series with a resistor, R_1, which limits the current drain on the blocking-bias source when the key is closed. R_2C_1 is a resistance-capacity filter that controls the rise time on make, the rise time increasing as $R_2 \times C_1$ is made larger. $C_1 \times (R_1 + R_2)$ controls the decay time on break in the same manner. Since grid current flows through R_2 in Fig. 8-4A when the key is closed, operating bias is developed, and R_2 is the normal grid leak for the tube. Thus C_1 only is varied to obtain the proper rise time.

With blocked-grid keying only a small current is broken compared with other systems, and sparking at the key is slight.

CATHODE KEYING

Keying the cathode circuit of a tube simultaneously opens the grid and plate circuits of the tube. This is shown in Fig. 8-5A. The condenser *C* serves as a short path for the r.f. energy, since the keying leads are often long. When a filament-type tube is keyed in this manner, the key is connected in the filament-transformer center-tap lead, as in Fig. 8-5B, and the system is called center-tap keying. The condensers *C* serve the same purpose as in cathode keying.

Cathode (or center-tap) keying results in less sparking at the key contacts than does plate-supply keying, for the same plate power. When used with an oscillator it does not respond as readily to key-click filtoring as does plate-circuit keying, but it is an excellent method for amplifier keying. If plate voltages above 300 are used, it is highly advisable to use a keying relay, to avoid accidental electrical shock at the key.

KEYING RELAYS

A keying relay can be substituted for a key in any of the keying circuits shown in this chapter. Most keying relays operate from 6.3 or 115 volts a.c., and they should be selected for their speed of operation and adequate insulation for the job to be done. Adequate current-handling capability is also a factor. A typical circuit is shown in Fig. 8-6.

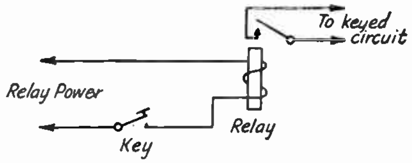


Fig. 8-6 — A keying relay can always be substituted for the key, to provide better isolation from the keyed circuit. An r.f. filter is generally required at the key, and the keying filter is connected in the keyed circuit at the relay contacts.

The relay-coil current that is broken by the key will cause clicks in the receiver, and an r.f. filter (see later in this chapter) is often necessary across the key. The normal keying filter connects at the relay armature contacts in the usual manner. Vibration effects of the keying relay upon the oscillator circuit should be avoided.

Key-Click Reduction

As pointed out earlier, interference caused by the key breaking current and the fast rise and decay times of the keyed characters is called "key clicks." The elimination of the interference depends upon its type.

R.F. FILTERS

Key clicks caused by the spark (often very minute) at the key contacts can be minimized by isolating the key from the rest of the wiring by a small r.f. filter. Such a filter usually consists of an r.f. choke in each key lead, placed right at the key terminals and by-passed on the line side by a small condenser. Such a circuit is shown in Fig. 8-7. Suitable values are best found by experiment, although 2.5-mh. r.f. chokes and a 0.001-μfd. condenser represent good starting points. The chokes must be capable of carrying the current that is broken, and the condenser must have a voltage rating equal at least to the voltage across the key under key-up conditions. Sometimes a small condenser directly across the key terminals is also necessary to remove the last trace of click. This type of r.f. filter is required in nearly every keying installation, in addition to the various circuits to be described in the following few paragraphs.

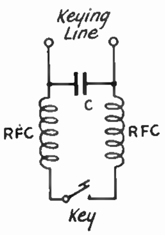


Fig. 8-7 — R.f. filter used for eliminating the radiation effects of sparking at the key contacts. Suitable values for best results with individual transmitters must be determined by experiment. Values for *RFC* range from 2.5 to 80 millihenrys and for *C* from 0.001 to 0.1 μfd.

Keying Filters

A filter used to give a desired shape to the keyed character, to eliminate clicks on the amateur bands and adjacent frequencies, is called a **keying filter** or **lag circuit.** In its simplest form it consists of a condenser and an inductance, connected as in Fig. 8-8. This type of keying filter is suitable for use in the circuits shown in Figs. 8-1 and 8-5. The optimum values of capacitance and inductance must be found by experiment but are not very critical. If a high-voltage low-current circuit is being keyed, a small condenser and a large inductance will be required, while a low-voltage high-current circuit needs a large condenser and small inductance to reduce the clicks properly. For example, a 300-volt 6-ma. circuit will require about 30 henrys and 0.05 μfd., while a 300-volt 50-ma. circuit needs about 1 henry and 0.5 μfd. For any given set of conditions, increasing

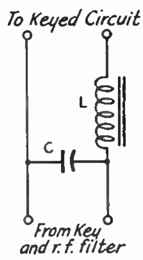


Fig. 8-8 — Lag circuit used for shaping the keying character to eliminate unnecessary sidebands. Actual values for any given circuit must be determined by experiment, and may range from 1 to 30 henrys for *L* and from 0.05 to 0.5 μfd. for *C*, depending on the keyed current and voltage.

the inductance will reduce the clicks on "make" and increasing the capacitance will reduce the clicks on "break."

Primary keying is adjusted by changing the filter values (*L* or *C* in Fig. 8-2). Since it is unlikely that a variety of chokes will be available to the operator, capacitance changes are usually all that can be made. If the keying is found too "soft," the value of *C* must be reduced.

Blocked-grid keying is adjusted by changing the values of resistors and condensers in the circuit, as outlined under the description of the circuit. The values required for individual installations will vary with the amount of blocking voltage and the value of grid leak.

Tube Keying

A tube keyer is a convenient device for keying the transmitter, because it allows the keying characteristic to be adjusted easily and also removes all dangerous voltages from the key itself. The current broken by the key is negligible and usually no r.f. filter is required at the key. A tube keyer uses a tube (or tubes in parallel) to control the current in the plate or cathode circuit of the stage being keyed. The keyer tube turns off the current flow when a high negative voltage is applied to the grid of the keyer tube. The keying characteristic is shaped through the time constants of the grid circuit of the keyer tube, in exactly the same way that it is controlled in blocked-grid keying. When a tube keyer is used to replace the key in a plate or cathode circuit, the power output of the stage may be reduced somewhat because the voltage drop across the keyer lowers the plate voltage or adds cathode bias, but this is of little importance and can be minimized by using more keyer tubes in parallel.

A Vacuum-Tube Keyer

A tube-keyer unit is shown in Figs. 8-9 and 8-10. T_1, the 80 rectifier, and C_1 and R_1 form the power-supply section that furnishes the blocking voltage for the keyer tubes. S_1 and S_2 and their associated resistors and condensers are included to allow the operator to select the keying characteristic he wants. A simplified version could omit the switches and extra components, since once the values have been selected the components can be soldered permanently in place. The rule for adjusting the keying characteristic is the same as for blocked-grid keying. However, large values of resistors and small values of condensers can be used, since there is no value of grid leak determined by the tube that dictates a starting point.

As many 45s may be added in parallel as desired. The voltage drop through a single tube varies from about 90 volts at 50 ma. to 50 volts at 20 ma. Tubes added in parallel will reduce the drop in proportion to the number of tubes used.

When connecting the output terminals of the keyer to the circuit to be keyed, the grounded output terminal of the keyer must be con-

Fig. 8-9 — A vacuum-tube keyer, built up on a 7 × 9 × 2-inch chassis with space for four or less keyer tubes and the power-supply rectifier. The resistors and condensers that produce the lag are underneath, controlled by the knobs at the right. The jack is for the key, while terminals at the left are for the keyed circuit.

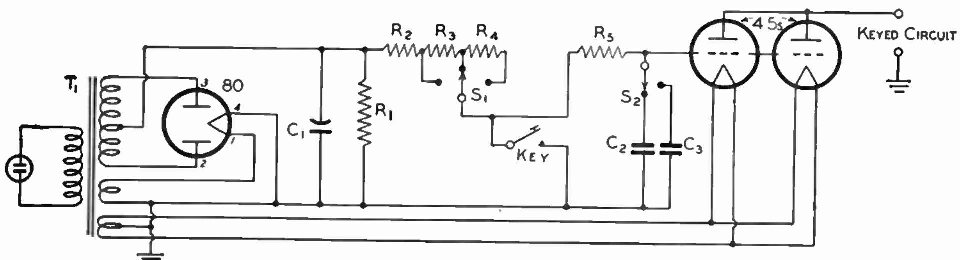


Fig. 8-10 — Wiring diagram of a practical vacuum-tube keyer similar to the one in Fig. 8-9.

C_1 — 2-μfd. 600-volt paper.
C_2 — 0.0033-μfd. mica.
C_3 — 0.0047-μfd. mica.
R_1 — 0.22 megohm, 1 watt.
R_2 — 50,000 ohms, 10 watts.
R_3, R_4 — 4.7 megohms, 1 watt.
R_5 — 0.47 megohm, 1 watt.
S_1, S_2 — 3-position 1-circuit rotary switch.
T_1 — 325-0-325 volts, 5 volts and 2.5 volts (Thordarson T-13R01).

nected to the transmitter ground. Thus the keyer can be used only in negative-lead or cathode keying.

When the key or keying lead has poor insulation, the resistance may become low enough (particularly in humid weather) to reduce the blocking voltage and allow the keyer tube to pass some current. This may cause a slight back-wave, but can be cured by better insulation or reduced values of R_2, R_3, R_4 and R_5.

Checking Transmitter Keying

One of the best ways to check your transmitter keying is to enlist the aid of a near-by amateur and trade stations with him for a short time. Not only will you be able to check your own key clicks and chirps, but if you have any complaint about the other fellow's signal this is a convenient way to let him know!

SIGNAL CHECKING

If keying the transmitter does not affect the line voltage, the station communications receiver can be used to check keying. The antenna should be disconnected from the receiver and the antenna posts shorted to ground. This method is satisfactory only when the line voltage is not affected by keying.

Key Clicks

When checking for key clicks, the b.f.o. and a.v.c. of the monitoring receiver should be turned off. The keying should be adjusted so that a slight click is heard as the key is closed but practically none heard as it is opened. The gain should be reduced during these tests, since false clicks can be generated if the receiver is overloaded. No clicks should be heard off the signal frequency. Checks should also be made with no r.f. power but with the key breaking its normal current, to see if local clicks are generated by sparking at the key.

Chirps

Keying chirps may be checked by tuning in the signal or one of its harmonics on the highest frequency range of the receiver or monitor and listening to the beat note in the normal manner. The gain should be sufficient to give moderate signal strength, but low enough to avoid overloading. Adjust the tuning to give a low-frequency beat note and key the transmitter at several different speeds. The signal should be tuned in on either side of zero beat and at various beat frequencies for a complete check. Listening to a harmonic magnifies the effect of any chirp by the order of the harmonic.

Oscillator Keying

Any oscillator, either crystal- or self-controlled, should oscillate at low voltages (two or three volts) and have negligible change in frequency with plate voltage, if it is to key without chirps or clicks. A crystal oscillator will oscillate at low voltages if a regenerative type such as the Tri-tet or grid-plate is used and if an r.f. choke is connected in series with the grid-leak resistor, to reduce loading on the crystal. Oscillators of this type are generally free from chirp unless the crystal is a poor one or if there is too much air gap in the holder.

Self-controlled oscillators are more difficult to operate without chirp, but the important requirements are proper C/L ratio in the tank circuit, low plate (and screen) currents, and careful adjustment of the feed-back. A keyed self-controlled oscillator should be designed for best keying rather than maximum output.

Stages Following Keying

When a keying filter is being adjusted, the stages following the keyed stage should be made inoperative by removing the plate voltage. The following stages should then be connected in, one at a time, and the keying checked after each addition. An increase in click intensity (for the same carrier strength in the receiver) indicates that the clicks are being added in the stages following the one being keyed. The fixed bias on such stages should be sufficient to reduce the idling plate current (no excitation) to a low value, but not to zero. The output condensers on the filters of the power supplies feeding these later stages should not be too small.

Low-frequency parasitic oscillations can cause key clicks removed from the signal frequency by 50 or 100 kc. They are most common in beam-tetrode stages, and often can only be eliminated by neutralizing the stage.

Fig. 8-11 — A top view of the "Monitone." The shaft of the screwdriver-adjusted potentiometer controlling tone and volume is located between the 6J5 and 6SL7 tubes. The right-hand switch controls the a.c. and the center switch cuts the tone oscillator in and out.

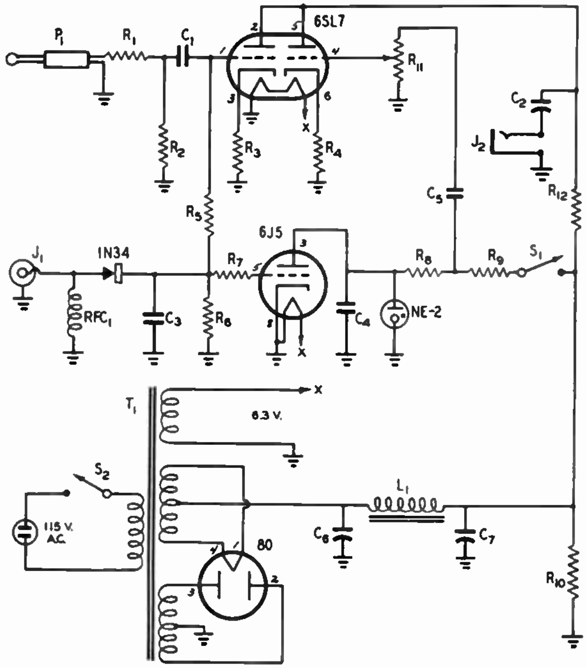


Fig. 8-12 — Circuit diagram of the "Monitone."

C_1 — 0.0047-μfd. 400-volt paper or mica.
C_2 — 0.1-μfd. 600-volt paper.
C_3 — 100-μμfd. mica.
C_4 — 0.001-μfd. paper or mica.
C_5 — 220-μμfd. paper or mica.
C_6, C_7 — 8-μfd. 450-volt electrolytic.
R_1 — 6800 ohms.
R_2 — 1000 ohms.
R_3, R_4 — 1200 ohms.
R_5 — 0.56 megohm.
R_6 — 1 megohm.
R_7 — 68,000 ohms.
R_8 — 4.7 megohms.
R_9 — 2.2 megohms.
R_{10} — 25,000 ohms, 5 watts.
R_{11} — 1-megohm volume control.
R_{12} — 22,000 ohms, 1 watt.
All resistors ½-watt unless otherwise specified.
L_1 — 40-ma. filter choke (Thordarson 13C26).
J_1 — Coaxial-connector jack.
J_2 — Open-circuit jack.
P_1 — 'Phone plug.
RFC_1 — 2.5-mh. r.f. choke.
S_1, S_2 — S.p.s.t. toggle.
T_1 — Small b.c. replacement transformer (Stancor P-6297).

MONITORING OF KEYING

The most popular type of monitor is an audio oscillator that is keyed simultaneously with the transmitter. A unit that will key automatically with the transmitter (and also blank the receiver output at the same time) is the "Monitone" shown in Figs. 8-11 and 8-13. It requires no direct connection to the transmitter or key. When the key is up, signals from the receiver pass through the Monitone to the headphones. When the key is down, the receiver output is blanked and a sidetone appears in the headphones. The sidetone and blanking are keyed by the r.f. output of the transmitter, regardless of frequency.

The wiring diagram of the Monitone is given in Fig. 8-12. The 6SL7 acts as a dual amplifier, for the receiver output and for the sidetone oscillator (consisting of the neon bulb — NE-2, C_4 and $R_8 + R_9$). When r.f. from the transmitter is fed in at J_1 it is rectified by the 1N34 and a negative voltage appears at the grid of the 6J5 and Pin 1 of the 6SL7. This negative voltage cuts off the 6J5 and the one half of the 6SL7. The neon-bulb oscillator goes into action and the resultant tone is amplified by the other half of the 6SL7. S_1 is opened for 'phone operation.

The arrangement of parts can be seen in Figs. 8-11 and 8-13. The placement is not critical, although the r.f. network (RFC_1, 1N34 and C_3) should be separated from the 6SL7.

Installation of the device consists of plugging the input lead into the headphone jack on the receiver, the headphones into the jack on the Monitone, and the plug into the a.c. outlet.

The final step is to couple the right amount of r.f. into the Monitone. A short piece of wire can be connected to the coaxial fitting on the back of the Monitone if the operating table is near the transmitter. If they are widely separated a piece of RG-59/U or ordinary shielded wire can be run from the coaxial fitting to a point near the final amplifier or feeders. (CAUTION — high voltage!) The length of the pick-up wire, either directly from the Monitone or extending beyond the shielding of the coaxial

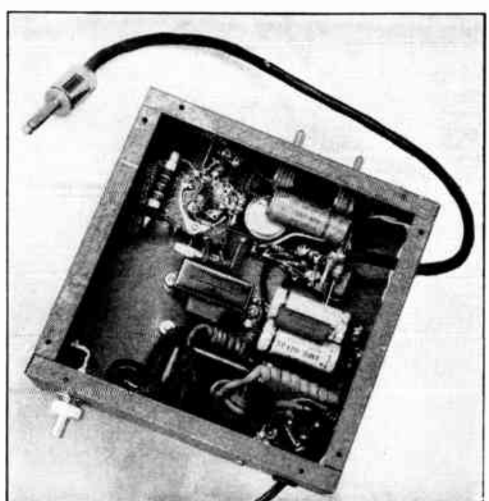


Fig. 8-13 — Bottom view of the "Monitone," showing the small neon lamp supported on its own leads and located directly above the center of the 6J5 socket. The r.f. feeder terminates on the tie-point that carries the 1N34.

line, will depend on the transmitter power being used. Only a foot or two will be needed.

Close the key and move the pick-up nearer or farther from the transmitter or feeder until the neon bulb in the Monitone glows. Find a point where a little less coupling will extinguish the neon — in other words adjust for the loosest coupling that will cause vigorous and sustained oscillation of the neon circuit. If only the final is keyed, care must be taken not to put the pick-up wire in the r.f. field of the driver stages — otherwise the oscillator will run continuously whenever the transmitter is switched on.

Break-In Operation

Break-in operation requires a separate receiving antenna, since none of the available antenna change-over relays is fast enough to follow keying. The receiving antenna should be installed as far as possible from the transmitting antenna. It should be mounted at right angles to the transmitting antenna and fed with low pick-up lead-in material such as coaxial cable or 300-ohm Twin-Lead, to minimize pick-up.

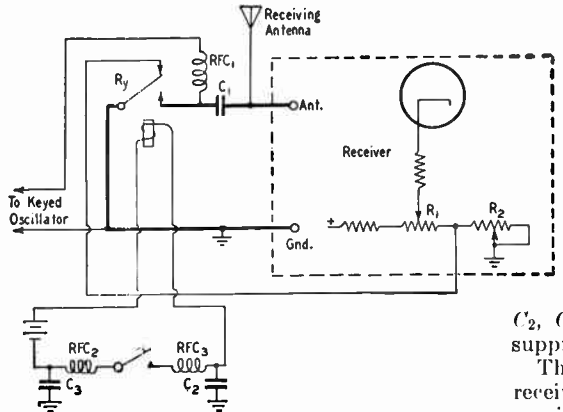


Fig. 8-14 — Wiring diagram for smooth break-in operation. The leads shown as heavy lines should be kept as short as possible, for minimum pick-up of the transmitter signal.
C_1, C_2, C_3 — 0.001 μfd.
R_1 — Receiver manual gain control.
R_2 — 5000- or 10,000-ohm wire-wound potentiometer.
RFC_1, RFC_2, RFC_3 — 2.5-mh. r.f. choke.
Ry — S.p.d.t. keying relay.

If a low-powered transmitter is used, it is often quite satisfactory to use no special equipment for break-in operation other than the separate receiving antenna, since the transmitter will not block the receiver too seriously. Even if the transmitter keys without clicks, some clicks will be heard when the receiver is tuned to the transmitter frequency because of overload in the receiver. An output limiter, as described in Chapter Five, will wash out these clicks and permit good break-in operation even on your transmitter frequency.

When powers above 25 or 50 watts are used, special treatment is required for quiet break-in on the transmitter frequency. A means should be provided for shorting the input of the receiver when the code characters are sent, and a means for reducing the gain of the receiver at the same time is often necessary. The system shown in Fig. 8-14 permits quiet break-in operation for higher-powered stations. It requires a simple operation on the receiver but otherwise is perfectly straightforward. R_1 is the regular receiver r.f. and i.f. gain control. The ground lead is lifted on this control and run to a rheostat, R_2, that goes to ground. A wire from the junction runs outside the receiver to the keying relay, R_y. When the key is up, the ground side of R_1 is connected to ground through the relay arm, and the receiver is in its normal operating condition. When the key is closed, the relay closes, which breaks the ground connection from R_1 and applies additional bias to the tubes in the receiver. This bias is controlled by R_2. When the relay closes, it also closes the circuit to the transmitter oscillator. C_2, C_3, RFC_2 and RFC_3 compose a filter to suppress the clicks caused by the relay current.

The keying relay should be mounted on the receiver as close to the antenna terminals as possible, and the leads shown heavy in the diagram should be kept short, since long leads will allow too much signal to get through into the receiver. A good high-speed keying relay should be used. If a two-wire line is used from the receiving antenna, another r.f. choke, RFC_4, will be required. The revised portion of the schematic is shown in Fig. 8-15.

A DE LUXE BREAK-IN SYSTEM

In many instances it is quite difficult to key an oscillator without clicks *and* chirps. Most oscillators will key without apparent chirp

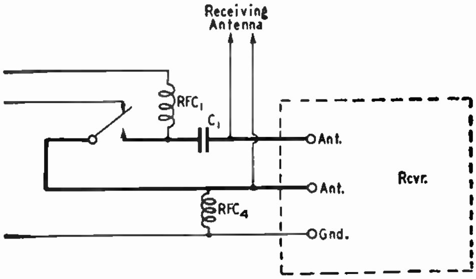


Fig. 8-15 — Necessary circuit revision of Fig. 8-14 if a two-wire lead from the receiving antenna is used. RFC_4 is a 2.5-mh. r.f. choke — other values are the same as in Fig. 8-14.

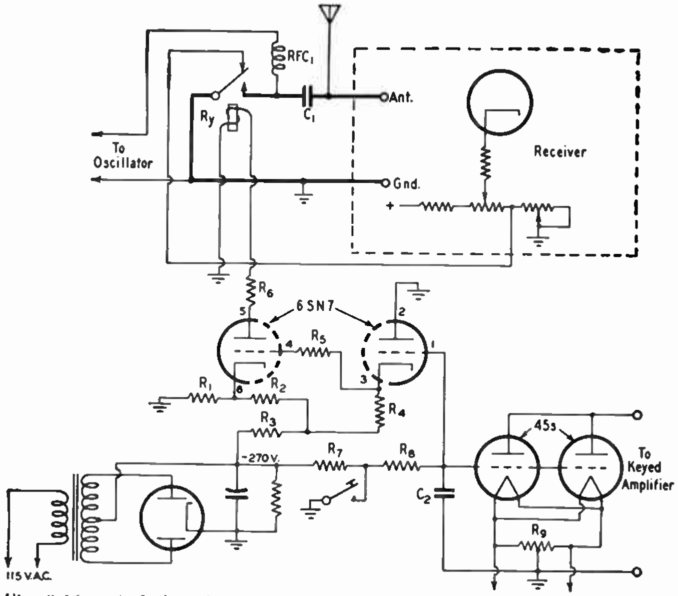


Fig. 8-16 — A de luxe break-in system that holds the oscillator circuit closed (and the receiver input shorted) during a string of fast dots but opens between letters or words.

C_1 — 0.001-μfd. mica.
C_2 — 0.0047-μfd. mica.
R_1 — 20,000 ohms, 10 watts, wire-wound.
R_2 — 1800 ohms.
R_3 — 1500 ohms.
R_4, R_5 — 1.0 megohm.
R_6 — 4700 ohms.
R_7 — 6.8 megohm.
R_8 — 0.47 megohm.
R_9 — 50-ohm center-tapped resistor, 2 watts.
All resistors 1-watt composition unless otherwise noted.
RFC_1 — 2.5-mh. r.f. choke.
Ry — High-speed relay, 1400-ohm 18-volt coil (Stevens-Arnold Type 172 Millisec relay).

if the rise and decay times are made very short, but this introduces key clicks that cannot be avoided. The system shown in Fig. 8-16 avoids this trouble by turning on the oscillator quickly, keying an amplifier with a vacuum-tube keyer, and turning off the oscillator after the amplifier keying is finished. The oscillator is turned on and off without lag, but the resultant clicks are not passed through the transmitter. Actually, with keying speeds faster than about 15 w.p.m., the oscillator will stay turned on for a letter or even a word, but it turns off between words and allows the transmitting station to hear the "break" signal of the other station. It requires one tube more than the ordinary vacuum-tube keyer and a special high-speed relay.

As can be seen from Fig. 8-16, the circuit is a combination of the break-in system of Fig. 8-14 and the tube keyer of Fig. 8-10, with a 6SN7 tube and a few resistors added. Normally the left-hand portion of the 6SN7 is biased to a low value of plate current by the drop through R_2 (part of the bleeder $R_1R_2R_3$) and the relay is open. When the key is closed and C_2 starts to discharge, the right-hand portion of the 6SN7 draws current and this in turn puts a less-negative voltage on the grid of the left-hand portion. The tube draws current and the relay closes. The relay will stay closed until the negative voltage across C_2 is close to the supply voltage, and consequently a string of dots or dashes (which doesn't give C_2 a chance to charge to full negative) will keep the relay closed. In adjusting the system, R_2 controls the amount of idling current through the relay and R_6 determines the voltage across the relay. R_7, R_8 and C_2 are the normal resistors and condenser for the tube keyer. When adjusted properly, the relay will close without delay on the first dot and open quickly during the spaces between words or slower letters. When idling, the voltage across the relay should be one or two volts — with the key down it should be 18 volts.

The oscillator should be designed to key as fast as possible, which means that series resistances and shunt capacitances should be held to a minimum. Negative plate-lead keying is slightly faster than cathode keying and should be used in the oscillator. The keyer tubes are connected in the cathode circuit of an amplifier stage, far enough removed in the circuit to avoid reaction on the oscillator.

● ELECTRONIC KEYS

Electronic keys, as contrasted with mechanical automatic keys, use vacuum tubes (and possibly relays) to form automatic dashes as well as automatic dots. Full descriptions of such devices can be found in the following *QST* articles:

Beecher, "Electronic Keying," April, 1940.
Grammer, "Inexpensive Electronic Key," May, 1940.
Savage, "Improved Switching Arrangement for Simplified Electronic Key," March, 1942.
Gardner, "New Electronic-Key Circuits," March, 1944.
Wiley, "Simplifying the Electronic Key," July, 1944.
"Electronic Bug Movement," Feb., 1945.
Snyder, "Versatile Electronic Key," March, 1945; correction, page 82, May, 1945.
Beecher, "Better Electronic Keyer," August, 1945.
DeHart, "De luxe Electronic Key," Sept., 1946; correction, page 27, Jan., 1947.
Gotisar, "The Dash Master," Aug., 1948.
Bartlett, "Further Advances in Electronic-Keyer Design," October, 1948; correction, page 10, Jan., 1949.

CHAPTER 9

Radiotelephony

To transmit intelligible speech by radio it is necessary to **modulate** the normally-constant output of the radio-frequency section of a transmitter. **Modulation,** defined in the most simple terms, is the process of varying the transmitter output in a desired fashion. In the case of radiotelephony, it means varying the radio-frequency output in a way that follows the spoken word.

The unmodulated r.f. output of the transmitter is called the **carrier**. In itself, the carrier conveys no information to the receiving operator — other than that the transmitting station is "on the air." It is only when the carrier is modulated that it becomes possible to transmit a message.

METHODS OF MODULATION

The carrier as generated by the transmitter is a simple form of alternating current — practically a sine wave. As such, it has three "dimensions" that can be varied — its amplitude, its frequency, and its phase. Modulation can be applied successfully to any of the three.

In **amplitude modulation (AM)** the amplitude of the carrier is made to vary upward and downward, following similar variations in audio-frequency currents generated by a microphone. In this type of modulation the frequency and phase of the carrier are unaffected by the modulation. Amplitude modulation is today the most widely-used system in amateur stations.

In **frequency modulation (FM)** the frequency of the carrier is made to vary above and below the unmodulated carrier frequency, the frequency variations being made to follow the a.f. currents. The power output of the transmitter does not change in frequency modulation. The *phase* of the carrier does change, however, since frequency and phase are intimately related.

In **phase modulation (PM)** the phase of the carrier is advanced and retarded by the modulating audio-frequency current. The transmitter power does not vary with modulation, but the carrier frequency changes.

These definitions are quite broad, and detailed explanations of the three systems are given later in this chapter.

SIDEBANDS

No matter what the method of modulation, the process of modulating a carrier sets up new groups of radio frequencies both above and below the frequency of the carrier itself. These new frequencies that accompany the modulation are called **side frequencies,** and the frequency bands occupied by a group of them when the modulating signal is complex (as it is with voice modulation) are called **sidebands.** Sidebands always appear on *both* sides of the carrier; the band higher than the carrier frequency is called the **upper sideband** and the band lower than the carrier frequency is called the **lower sideband**. The modulation (that is, the intelligence) in the signal is carried in the sidebands, not in the carrier itself.

The result of this is that a modulated signal occupies a group or band of frequencies **(channel)** rather than the single frequency of the carrier alone. Just how much of a frequency band (that is, how wide a channel) is occupied depends upon the method of modulation and the frequency characteristics of the modulating signal itself.

A normal voice contains frequencies or tones ranging from perhaps a hundred cycles at the low end to several thousand cycles at the high end. Vowel sounds (*a*, *e*, *i*, *o*, *u*) are in general fairly low in frequency and contain most of the voice power. Consonants usually are characterized by higher frequencies, and the hissing sound of the letter "S" is particularly high up in the audio-frequency range. The timbre of a voice, or the thing that makes it possible for us to distinguish the voices of different individuals, results principally from overtones or harmonics. All these things add up to the fact that a fairly wide range of audio frequencies is needed for the accurate reproduction of a *particular* voice.

On the other hand, the frequency range required for good *intelligibility* is not nearly so wide as that needed for accurate reproduction or "fidelity." For the latter, an audio system that is "flat" — that is, has the same amplification at all frequencies — over the range up to about 10,000 cycles is required. But a system that "cuts off" above 2500 cycles — that is, has comparatively little output above that figure — will transmit everything that is necessary for *understandable* speech. The speech may sound a little less like the speaker's actual voice, but it will be thoroughly intelligible to the receiving operator.

This distinction between intelligibility and "quality" is extremely important. The *minimum* channel occupied by a 'phone transmitter, *no matter what the system of modulation*, is equal to *twice the highest audio frequency present in the modulation.* If audio frequencies up to 10,000 cycles are contained in the modulation, the channel will be at least twice 10,000 or 20,000 cycles (20 kc.) wide. But if there are no frequencies above 2500 cycles in the modulation, the channel will be only 5000 cycles (5 kc.) wide. In amateur bands where there is a great deal of congestion it is in everybody's interest that each transmitter should occupy no more than the minimum channel needed for transmitting *intelligible* speech. Taking up a wider frequency channel than that simply creates unnecessary interference.

Also, transmitting a wide range of audio frequencies in a congested band actually accomplishes nothing, insofar as "fidelity" is concerned; the receiving operator has to use so much receiver selectivity — in order to "copy" the signal at all — that the higher-frequency sidebands are rejected by the receiver. Those sidebands do, however, continue to interfere with stations operating on near-by carrier frequencies.

We have said that the *minimum* channel is equal to twice the highest audio frequency in the modulation. The actual channel occupied may be several times the minimum necessary channel-width. This depends on the system of modulation used, for one thing. For another, it depends on whether the system is operated properly or whether it is misadjusted. Improper operation of any sort invariably increases the channel-width. Since the amount of frequency space available for amateur operation is limited, no operator of an amateur 'phone station can avoid the obligation to confine his transmissions to the least possible space.

Amplitude Modulation

In amplitude modulation, as we have already stated, the amplitude or strength of the carrier is varied up and down from the unmodulated value. The several methods of making the carrier strength vary are discussed in a later section; for the moment let us look only at the end result that is the object of all the various amplitude-modulation systems.

In Fig. 9-1, the drawing at A shows the unmodulated r.f. carrier, assumed to be a sine wave of the desired radio frequency. The graph can be taken to represent either voltage or current, and each cycle has just the same height as the preceding and following ones.

In B, the carrier wave is assumed to be modulated by a signal having the shape shown in the small drawing above. The frequency of the modulating signal is much lower than the carrier frequency, so quite a large number of carrier cycles can occur during each cycle of the modulating signal. This is a necessary condition for good modulation, and always is the case in radiotelephony because the audio frequencies used are very low compared with the radio frequency of the carrier. (Actually, there would be very many times more r.f. cycles in each modulation cycle than are shown in the drawing; so many that it is impossible to make the drawing to actual scale.) When the modulating signal is "positive" (above its axis) the carrier amplitude is increased *above* its unmodulated amplitude; when the modulating signal is "negative" the carrier amplitude is *decreased.* Thus the carrier grows larger and smaller with the polarity and amplitude of the modulating signal.

The drawing at C shows what happens with a stronger modulating signal. In this case the strength of the modulation is such that on the "up" modulation the carrier amplitude is doubled at the instant the modulating signal reaches its positive peak. On the negative peak of the modulating signal the carrier amplitude just reaches zero; in other words, the carrier is "all used up."

Percentage of Modulation

When a modulated wave is detected in a receiver the sound that comes out of the loud-

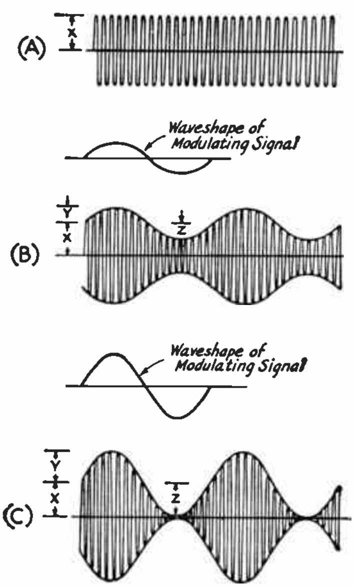


Fig. 9-1 — Graphical representation of (A) carrier unmodulated, (B) modulated 50%, (C) modulated 100%.

speaker or headset is caused by the modulation, not by the carrier. In other words, in detecting the signal the receiver eliminates the carrier and takes from it the modulating signal. The stronger the modulation, therefore, the greater is the useful receiver output. Obviously, it is desirable to make the modulation as strong or "heavy" as possible. A wave modulated as in Fig. 9-1C would produce considerably more useful signal than the one shown at B.

The "depth" of the modulation is expressed as a percentage of the unmodulated carrier amplitude. In either B or C, Fig. 9-1, X represents the unmodulated carrier amplitude, Y is the maximum *increase* in amplitude on the modulation up-peak, and Z is the maximum *decrease* in amplitude on the modulation down-peak. Assuming that Y and Z are equal, then the **percentage of modulation** can be found by dividing either Y or Z by X and multiplying the result by 100. In the wave shown in Fig. 9-1C, Y and Z are both equal to X, so the wave is modulated 100 per cent. In case the modulation is not symmetrical (Y and Z not equal), the larger of the two should be used for calculating the percentage of modulation.

The outline of the modulated wave is called the **modulation envelope.** It is shown by the thin line outlining the patterns in Figs. 9-1 and 9-2.

Power in Modulated Wave

The amplitude values shown in Fig. 9-1 correspond to current or voltage, so the drawings may be taken to represent instantaneous values of either. Now power varies as the square of either the current or voltage (so long as the resistance in the circuit is unchanged), so at the peak of the modulation up-swing the instantaneous power in the wave of Fig. 9-1C is four times the unmodulated carrier power (because the current and voltage are doubled). At the peak of the down-swing the power is zero, since the amplitude is zero. With a sine-wave modulating signal, the *average* power in a 100-per-cent modulated wave is one and one-half times the value of unmodulated carrier power; that is, the power output of the transmitter increases 50 per cent with 100-per-cent modulation.

The complex waveform of speech does not contain as much power as there is in a pure tone or sine wave of the same peak amplitude. On the average, speech waveforms will contain only about half as much power as a sine wave, both having the same peak amplitude. The average power output of the transmitter therefore increases only about 25 per cent with 100-per-cent speech modulation. However, the *instantaneous* power output must quadruple on the peak of 100-per-cent modulation regardless of the modulating waveform. Therefore, the peak output-power capacity of the transmitter must be the same for any type of modulating signal.

Overmodulation

If the carrier is modulated more than 100 per cent, a condition such as is shown in Fig. 9-2 occurs. Not only does the peak amplitude exceed twice the carrier amplitude, but there actually may be a considerable period during which the output is entirely cut off. Therefore the modulated wave is distorted, and the modulation contains harmonics of the audio modulating frequency.

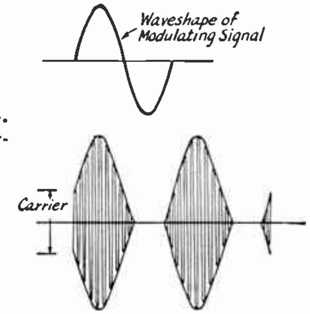


Fig. 9-2 — An overmodulated r.f. carrier wave.

The sharp "break" when the carrier is suddenly cut off on the modulation down-swing produces a type of distortion that contains a large number of harmonics. For example, it is easily possible for harmonics up to the fifth to be produced by a relatively small amount of overmodulation. If the modulating frequency is 2000 cycles, this means that the actual modulated wave will have sidebands not only at 2000 cycles, but also at 4000, 6000, 8000 and 10,000 cycles each side of the carrier frequency. The signal thus occupies five times the needed channel-width. It is obviously of first importance to prevent the modulation from exceeding 100 per cent, and thus prevent the generation of spurious sidebands — commonly called "splatter."

Carrier Requirements

For satisfactory amplitude modulation, the carrier *frequency* should be entirely unaffected by the application of modulation. If modulating the amplitude of the carrier also causes a change in the carrier frequency, the frequency will wobble back and forth with the modulation. This causes distortion and widens the channel taken by the signal. Thus unnecessary interference is caused to other transmissions. In practice, this undesirable frequency modulation is prevented by applying the modulation to an r.f. amplifier stage that is isolated from the frequency-controlling oscillator by a **buffer amplifier.** Amplitude modulation applied directly to an oscillator always is accompanied by frequency modulation. Under existing regulations amplitude modulation of an oscillator is permitted only on frequencies above 144 Mc. Below that frequency the regulations require that an amplitude-modulated transmitter be completely free from frequency modulation.

Plate Power Supply

The d.c. power supply for the plate or plates of the modulated amplifier must be well filtered; if it is not, the plate-supply ripple will modulate the carrier and cause annoying hum. To be substantially hum-free, the ripple voltage should not be more than about 1 per cent of the d.c. output voltage.

In amplitude modulation the plate current varies at an audio-frequency rate; in other words, an alternating current is superimposed on the d.c. plate current. The output filter condenser in the plate supply must have low reactance, at the lowest audio frequency in the modulation, if the transmitter is to modulate equally well at all audio frequencies. The condenser capacitance required depends on the ratio of d.c. plate current to plate voltage in the modulated amplifier. The requirements will be met satisfactorily if the capacitance of the output condenser is at least equal to

$$C = 25 \frac{I}{E}$$

where C = Capacitance of output condenser in μfd.

I = D.c. plate current of modulated amplifier in milliamperes

E = Plate voltage of modulated amplifier

Example: A modulated amplifier operates at 1250 volts and 275 ma. The capacitance of the output condenser in the plate-supply filter should be at least

$$C = 25 \frac{I}{E} = 25 \times \frac{275}{1250} = 25 \times 0.22 = 5.5 \mu\text{fd.}$$

Linearity

Up to the limit of 100-per-cent modulation, the amplitude of the r.f. output should be directly proportional to the amplitude of the modulating signal. Fig. 9-3 is a graph of an ideal **modulation characteristic,** or curve showing the relationship between r.f. output amplitude and modulating-signal amplitude. The modulation swings the amplitude back and forth along the curve A as the modulating signal alternately swings positive and negative. Assuming that the negative peak of the modulating signal is just sufficient to reduce the carrier amplitude to zero (modulating signal equal to -1 in the drawing), the same modulating signal peak in the *positive* direction ($+1$) should cause the r.f. amplitude to reach twice its unmodulated-carrier value. The ideal modulation characteristic is a straight line, as shown by curve A. Such a modulation characteristic is perfectly **linear.**

A **nonlinear** characteristic is shown by curve B. The r.f. amplitude does not reach twice the unmodulated carrier amplitude when the modulating signal reaches its positive peak. A modulation characteristic of this type gives a modulation envelope that is "flattened" on the up-peak; in other words, the modulation envelope is not an exact reproduction of the modulating signal. It is therefore distorted and harmonics are generated, causing the transmitted signal to occupy a wider channel than is necessary. A nonlinear modulation characteristic can easily result when a transmitter is not properly designed or is misadjusted.

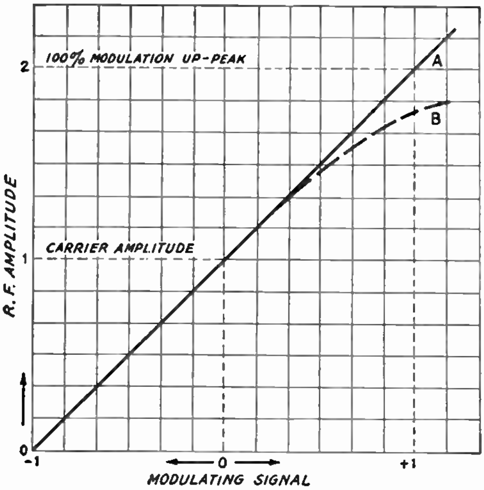


Fig. 9-3 — The modulation characteristic shows the relationship between the instantaneous amplitude of the r.f. output and the instantaneous amplitude of the modulating signal. The ideal characteristic is a straight line, as shown by curve *A*.

The **modulation capability** of the transmitter is the maximum percentage of modulation that is possible without objectionable distortion from nonlinearity. The maximum capability is 100 per cent on the down-peak but can be higher on the up-peak. The modulation capability should be as high as possible, so that the most effective signal can be transmitted for a given carrier power.

Types of Amplitude Modulation

The most widely-used amplitude-modulation system is that in which the modulating signal is applied in the plate circuit of a radio-frequency power amplifier **(plate modulation).** In a second type the audio signal is applied to a control grid **(grid-bias modulation).** A third system, involving variation of both plate and grid voltages, is called **cathode modulation.**

PLATE MODULATION

The most popular system of amplitude modulation is plate modulation. It is the simplest to apply, gives the highest efficiency in the modulated amplifier, and is the easiest to adjust for proper operation.

Fig. 9-4 shows the most widely-used system of plate modulation. A balanced (push-pull Class A, Class AB or Class B) **modulator** is transformer-coupled to the plate circuit of the modulated r.f. amplifier. The audio-frequency power generated by the modulator is combined with the d.c. power in the modulated-amplifier plate circuit by transfer through the coupling transformer, *T*. For 100-per-cent modulation the audio-frequency output of the modulator and the turns ratio of the coupling transformer must be such that the voltage at the plate of the modulated amplifier varies between zero and twice the d.c. operating plate voltage, thus causing corresponding variations in the amplitude of the r.f. output.

As stated earlier, the average power output of the modulated stage must increase during modulation. The modulator must be capable of supplying to the modulated r.f. stage sine-wave audio power equal to 50 per cent of the d.c. plate input. For example, if the d.c. plate power input to the r.f. stage is 100 watts, the sine-wave audio power output of the modulator must be 50 watts.

Modulating Impedance; Linearity

The **modulating impedance**, or load resistance presented to the modulator by the modulated r.f. amplifier, is equal to

$$\frac{E_b}{I_p} \times 1000$$

where E_b = D.c. plate voltage
I_p = D.c. plate current (ma.)

E_b and I_p are measured without modulation.

The power output of the r.f. amplifier must vary as the square of the plate voltage (the r.f. voltage must be proportional to the applied plate voltage) in order for the modulation to be linear. This will be the case when the amplifier operates under Class C conditions. The linearity then depends upon having sufficient grid excitation and proper bias, and upon the adjustment of circuit constants to the proper values.

Adjustment of Plate-Modulated Amplifiers

The general operating conditions for Class C operation have been described in Chapter Six. The grid bias and grid current required for plate modulation usually are given in the operating data supplied by the tube manufacturer; in general, the bias should be such as to give an operating angle of about 120 degrees at carrier plate voltage, and the grid excitation should be great enough so that the amplifier's plate efficiency will stay constant when the plate voltage is varied over the range from zero to twice the unmodulated value. For best linearity, the grid bias should be obtained partly from a fixed source of about the cut-off value, and then supplemented by grid-leak bias to supply the remainder of the required operating bias.

The maximum permissible d.c. plate power input for 100-per-cent modulation is twice the sine-wave audio-frequency power output of the modulator. This input is obtained by varying the loading on the amplifier (keeping its tank circuit tuned to resonance) until the product of d.c. plate voltage and plate current is the desired power. The modulating impedance under these conditions must be transformed to the proper value for the modulator by using the correct output-transformer turns ratio. This point is considered in detail later in this chapter in the section on Class B modulator design.

Neutralization, when triodes are used, should be as nearly perfect as possible, since regeneration may cause nonlinearity. The amplifier also must be completely free from parasitic oscillations.

Although the *effective* value of power input increases with modulation, as described above, the *average* d.c. plate power input to a plate-modulated amplifier does not change. This is because each increase in plate voltage and plate current is balanced by an equivalent decrease in voltage and current on the next half-cycle of the modulating signal. The d.c. plate current to a properly-modulated amplifier is always constant, with or without modulation. On the other hand, an r.f. ammeter connected in the antenna or transmission line will show an increase in r.f. current with modulation.

Screen-Grid Amplifiers

Screen-grid tubes of the pentode or beam-tetrode type can be used as Class C plate-modulated amplifiers by applying the modulation to both the plate and screen grid. The usual method of feeding the screen grid with the necessary d.c. and modulation voltage is shown in Fig. 9-5. The dropping resistor, *R*, should be of the proper value to apply normal d.c. voltage to the screen under steady carrier

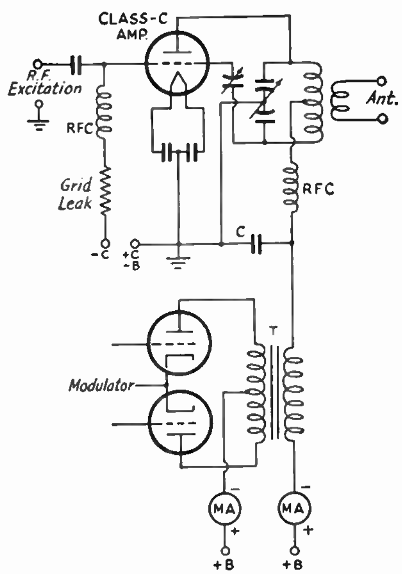


Fig. 9-4 — Plate modulation of a Class C r.f. amplifier. The r.f. plate by-pass condenser, C, in the amplifier stage should have reasonably high reactance at audio frequencies. (See section on Class B modulators.)

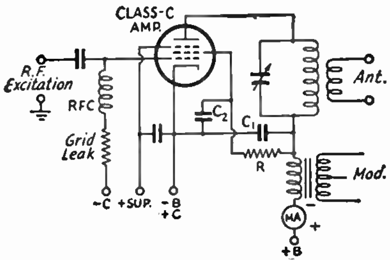


Fig. 9-5 — Plate and screen modulation of a Class C r.f. amplifier using a pentode tube. The plate r.f. by-pass condenser, C_1, should have reasonably high reactance at all audio frequencies. (See section on Class B modulators.) The screen by-pass, C_2, should be 0.002 μfd. or less in the usual case.

conditions. Its value can be calculated by taking the difference between plate and screen voltages and dividing it by the rated screen current.

The modulating impedance is found by dividing the d.c. plate voltage by the sum of the plate and screen currents. The plate voltage multiplied by the sum of the two currents gives the power input to be used as the basis for determining the audio power required from the modulator.

Modulation of the screen along with the plate is necessary because both elements affect the plate current in a power-type screen-grid tube, and a linear modulation characteristic cannot be obtained by modulating the plate alone. However, at least some types of beam tetrodes (the 4-250A and 4-125A, for example) can be modulated satisfactorily by applying the modulating power to the plate circuit alone, *provided* the screen is "floating" at audio frequencies — that is, is not grounded for a.f. but is connected to its d.c. supply through an audio impedance. The circuit is shown in Fig. 9-6. The choke coil L_1 is the audio impedance in the screen circuit; its inductance should be large enough to have a reactance (at the lowest desired audio frequency) that is not less than the impedance of the screen. The latter can be taken to be approximately equal to the d.c. screen voltage divided by the d.c. screen current.

Choke Coupling

Fig. 9-7 shows the circuit of the choke-coupled system of plate modulation. The d.c. plate power for both the modulator tube and modulated amplifier is furnished from a common source through the modulation choke, L. This choke must have high impedance, compared to the modulating impedance of the Class C amplifier, for audio frequencies. The modulator operates as a power amplifier with the plate circuit of the r.f. amplifier as its load, the audio output of the modulator being superimposed on the d.c. power supplied to the amplifier.

For 100-per-cent modulation, the audio voltage applied to the r.f. amplifier plate circuit across the choke, L, must have a peak value equal to the d.c. voltage on the modulated amplifier. To obtain this without distortion the r.f. amplifier must be operated at a *lower* d.c. plate voltage than the modulator. The extent of the voltage difference is determined by the type of modulator tube used. The necessary drop in voltage is provided by the resistor, R_1, which is by-passed for audio frequencies by the by-pass condenser, C_1.

This type of modulation seldom is used except in very low-power portable sets, because a Class A modulator is required. The output of a Class A modulator is very low compared to the power obtainable from a pair of tubes of the same size operated Class B, so only a small amount of r.f. power can be modulated.

GRID-BIAS MODULATION

Fig. 9-8 is the diagram of a typical arrangement for grid-bias modulation. In this system, the secondary of an audio-frequency output transformer, the primary of which is connected in the plate circuit of the modulator tube, is connected in series with the grid-bias supply for the modulated amplifier. The audio voltage varies the grid bias, and thereby the power output of the r.f. stage. The r.f. stage is operated as a Class C amplifier.

In this system the plate voltage is constant, and the increase in power output with modulation is obtained by making both the plate current and plate efficiency vary with the mod-

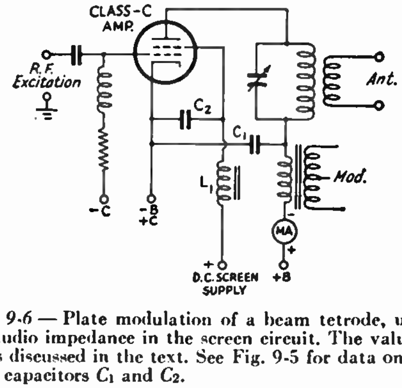


Fig. 9-6 — Plate modulation of a beam tetrode, using an audio impedance in the screen circuit. The value of L_1 is discussed in the text. See Fig. 9-5 for data on by-pass capacitors C_1 and C_2.

ulating signal. For 100-per-cent modulation, both plate current and efficiency must, at the peak of the modulation up-swing, be twice their carrier values. Thus at the modulation peak the power input is doubled, and since the plate efficiency also is doubled at the same instant the peak output power will be four times the carrier power. The maximum efficiency obtainable in practicable circuits is of the order of 70 to 80 per cent, so the carrier efficiency ordinarily cannot exceed about 35

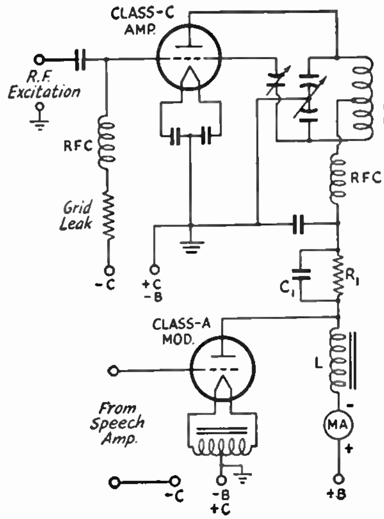


Fig. 9-7 — Choke-coupled plate modulation.

to 40 per cent. For a given r.f. tube, the carrier output is about one-fourth the power obtainable from the same tube plate-modulated.

Modulator Power

The increase in average carrier power with modulation is secured by varying the plate efficiency and d.c. plate input of the amplifier, so the modulator need supply only such power losses as may be occasioned by connecting it in the grid circuit. Since these are quite small, a modulator capable of only a few watts output will suffice.

The load on the modulator varies over the a.f. cycle as the rectified grid current of the modulated amplifier changes, so the modulator must have good voltage regulation. The purpose of the resistor *R* across the primary of the modulation transformer in Fig. 9-8 is to "swamp" such load changes by dissipating most of the audio power in the resistor. Generally, this resistor should be approximately equal to the load resistance required by the particular type of modulator tube used. The turns ratio of transformer *T* should be about 1-to-1 in most practical cases.

Grid-Bias Source

The change in instantaneous bias voltage with modulation causes the rectified grid current of the amplifier also to vary, the r.f. excitation being fixed. If the bias source has appreciable resistance, the change in grid current will cause a change in bias in a direction opposite to that caused by the modulation. It is necessary, therefore, to use a grid-bias source having low resistance, so that these bias variations will be negligible. Battery bias is satisfactory. If a rectified a.c. bias supply is used, the type having regulated output should be chosen (see Chapter Seven). Grid-leak bias for a grid-modulated amplifier is unsatisfactory, and its use should never be attempted.

Driver Regulation

The load on the r.f. driving stage varies with modulation, and a linear modulation characteristic cannot be obtained if the r.f. voltage from the driver does not stay constant with changes in load. Driver regulation (ability to maintain constant output voltage with changes in load) may be improved by using a driving stage having two or three times the power output necessary for excitation of the amplifier (which is less than the power required for ordinary Class C operation), and dissipating the extra power in a constant load such as a resistor. The variations caused by changes in load with modulation are thereby reduced because the variable load is only a fraction of the total load.

Operating Conditions

The d.c. plate input to the modulated amplifier, assuming a round figure of ⅓ (33 per cent) for the plate efficiency, should not exceed 1½ times the plate dissipation rating of the tube or tubes used in the modulated stage. On the modulation up-peaks the d.c. plate current doubles instantaneously but the d.c. plate voltage does not change. The problem, therefore, is to choose a set of operating conditions that will give normal Class C efficiency when the plate current is *twice* the carrier value.

Example: Two tubes having plate dissipation ratings of 55 watts each are to be used with grid-bias modulation. With plate modulation, the ratings are 1250 volts and 250 ma. for the two tubes, so the plate-voltage/plate-current ratio is

$$\frac{E\ \text{(volts)}}{I\ \text{(ma.)}} = \frac{1250}{250} = 5$$

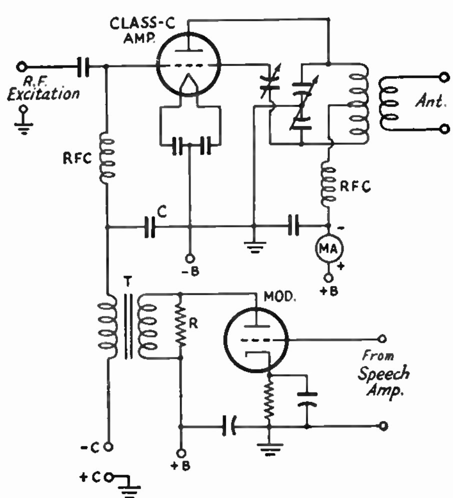


Fig. 9-8 — Grid-bias modulation of a Class C amplifier. The r.f. grid by-pass condenser, *C*, should have high reactance at audio frequencies (0.005 μfd. or less).

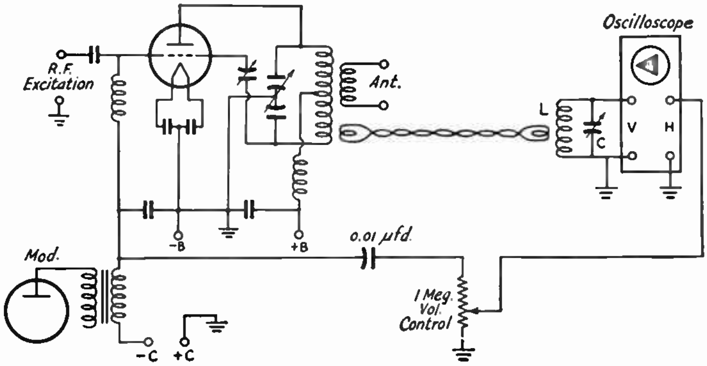


Fig. 9-9 — Adjustment set-up for grid-bias modulation. *L* and *C* should tune to the operating frequency, and may be coupled to the transmitter tank circuit through a twisted pair or other low-impedance line, using single-turn links at each end. The 0.01-μfd. blocking condenser that couples the audio voltage to the horizontal plates of the oscilloscope should have a voltage rating equal to about three times the grid bias.

With grid-bias modulation the maximum power input, at 33% efficiency, is

$P = 1.5 \times (2 \times 55) = 1.5 \times 110 = 165$ watts

The maximum recommended plate voltage for these tubes is 1500 volts. Using this figure, the plate current for the two tubes will be

$$I = \frac{P}{E} = \frac{165}{1500} = 0.11 \text{ amp.} = 110 \text{ ma.}$$

The plate-voltage/plate-current ratio at *twice* carrier plate current is

$$\frac{1500}{220} = 6.8$$

This is quite satisfactory. In this case it would be possible to use a lower plate voltage without having the plate-voltage/plate-current ratio drop to too low a value. At 1300 volts, for example, the ratio would be slightly over 5. However, at 1000 volts it would be only 3.

At 33% efficiency, the carrier output to be expected is 55 watts.

The tank-circuit *L/C* ratio should be chosen on the basis of *twice* the carrier plate current. In the example above, it would be based on a plate-voltage/plate-current ratio of 6.8. Note that if *carrier* conditions are used the ratio is 13.6, and a tank *L/C* ratio based on this figure would have a *Q* much too low for good coupling to the output circuit.

Since the amplifier operates in normal Class C fashion on the modulation up-peaks, the grid bias should be chosen for Class C operation at the plate voltage used. It may be higher if desired, but should never be lower. It is convenient to have an adjustable bias source for arriving at optimum operating conditions.

Adjustment

This type of modulated amplifier should be adjusted with the aid of an oscilloscope. The oscilloscope should be connected as shown in Fig. 9-9. A tone source for modulating the transmitter is a convenience, since a steady tone will give a steady pattern on the oscilloscope. A steady pattern is easier to study than one that flickers with voice modulation.

Having determined the permissible carrier plate current as previously described, apply r.f. excitation and plate voltage and, without modulation, adjust the plate loading to give the required plate current (keeping the plate tank circuit tuned to resonance). Next, apply modulation and increase the modulating signal until the modulation characteristic shows curvature (see later section in this chapter for use of the oscilloscope). If curvature occurs well below 100-per-cent modulation, the plate efficiency is too high. Increase the plate loading slightly and reduce the excitation to maintain the same plate current; then apply modulation and check the characteristic again. Continue this process until the characteristic is linear from the horizontal axis to twice the carrier amplitude. It is usually easier to obtain a more linear characteristic with high plate voltage and low current (carrier conditions) than with relatively low plate voltage and high plate current.

Suppressor Modulation

The circuit arrangement for suppressor-grid modulation of a pentode tube is shown in Fig. 9-10. The operating principles are the same as for grid-bias modulation. However, the r.f. excitation and modulating signals are applied to separate grids; this makes the system somewhat simpler to operate because best adjustment for proper excitation requirements and proper modulating-circuit requirements are more or less independent. The carrier plate efficiency is approximately the same as for grid-bias modulation, and the modulator power requirements are similarly small. With tubes having suitable suppressor-grid charac-

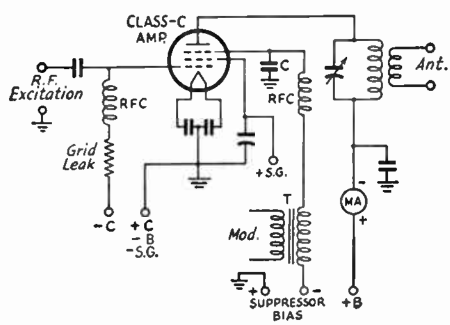


Fig. 9-10 — Suppressor-grid modulation of an r.f. amplifier using a pentode-type tube. The suppressor-grid r.f. by-pass condenser, *C*, should be the same as the grid by-pass condenser in grid-bias modulation (Fig. 9-8).

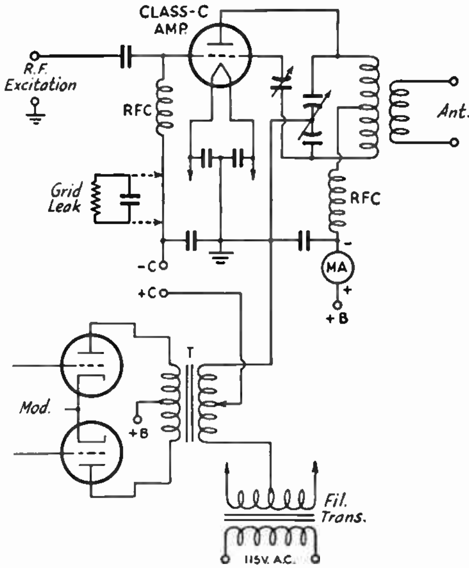


Fig. 9-11 — Circuit arrangement for cathode modulation of a Class C r.f. amplifier.

teristics, linear modulation up to practically 100 per cent can be obtained with negligible distortion.

The method of adjustment of this system is essentially the same as that described in the preceding paragraph.

CATHODE MODULATION

Circuit

The fundamental circuit for cathode or "center-tap" modulation is shown in Fig. 9-11. This type of modulation is a combination of the plate and grid-bias methods, and permits a carrier efficiency midway between the two. The audio power is introduced in the cathode circuit, and both grid bias and plate voltage vary during modulation.

Because part of the modulation is by the grid-bias method, the plate efficiency of the modulated amplifier must vary during modulation. The carrier efficiency therefore must be lower than the efficiency at the modulation peak. The required reduction in efficiency depends upon the proportion of grid modulation to plate modulation; the higher the percentage of plate modulation, the higher the permissible carrier efficiency, and vice versa. The audio power required from the modulator also varies with the percentage of plate modulation, being greater as this percentage is increased.

The way in which the various quantities vary is illustrated by the curves of Fig. 9-12. In these curves the performance of the cathode-modulated r.f. amplifier is plotted in terms of the tube ratings for plate-modulated telephony, with the percentage of plate modulation as a base. As the percentage of plate modulation is decreased, it is assumed that the grid-bias modulation is increased to make the over-all percentage of modulation reach 100 per cent. The limiting condition, 100-per-cent plate modulation and no grid-bias modulation, is at the right (A); pure grid-bias modulation is represented by the left-hand ordinate (B and C).

Example: Assume that the r.f. tube to be used has a 100% plate-modulation rating of 250 watts input and will give a carrier power output of 190 watts at that input. Cathode modulation with 40% plate modulation is to be used. From Fig. 9-12, the carrier efficiency will be 56% with 40% plate modulation, the permissible d.c. input will be 65% of the plate-modulation rating, and the r.f. output will be 48% of the plate-modulation rating. That is,

Power input $= 250 \times 0.65 = 162.5$ watts
Power output $= 190 \times 0.48 = 91.2$ watts

The required audio power, from the chart, is equal to 20% of the d.c. input to the modulated amplifier. Therefore

Audio power $= 162.5 \times 0.2 = 32.5$ watts

The modulator should supply a small amount of extra power to take care of losses in the grid circuit. These should not exceed four or five watts.

Modulating Impedance

The modulating impedance of a cathode-modulated amplifier is approximately equal to

$$m\frac{E_b}{I_b}$$

where m = Percentage of plate modulation (expressed as a decimal)
E_b = D.c. plate voltage on modulated amplifier
I_b = D.c. plate current of modulated amplifier

Example: Assume that the modulated amplifier in the example above is to operate at a plate

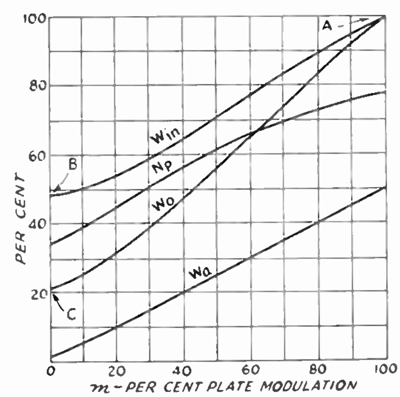


Fig. 9-12 — Cathode-modulation performance curves, in terms of percentage of plate modulation plotted against percentage of Class C telephony tube ratings.
W_{in} — D.c. plate input watts in terms of percentage of plate-modulation rating.
W_o — Carrier output watts in per cent of plate-modulation rating (based on plate efficiency of 77.5%).
W_a — Audio power in per cent of d.c. watts input.
N_p — Plate efficiency of the amplifier in percentage.

potential of 1250 volts. Then the d.c. plate current is

$$I = \frac{P}{E} = \frac{162.5}{1250} = 0.13 \text{ amp. (130 ma.)}$$

The modulating impedance is

$$m\frac{E_b}{I_b} = 0.4\frac{1250}{0.13} = 3846 \text{ ohms}$$

The modulating impedance is the load into which the modulator must work, just as in the case of pure plate modulation. This load must be matched to the load required by the modulator tubes by proper choice of the turns ratio of the modulation transformer.

Conditions for Linearity

R.f. excitation requirements for the cathode-modulated amplifier are midway between those for plate modulation and grid-bias modulation. More excitation is required as the percentage of plate modulation is increased. Grid bias should be considerably beyond cut-off; fixed bias from a supply having good voltage regulation is preferred, especially when the percentage of plate modulation is small and the amplifier is operating more nearly like a grid-bias modulated stage. At the higher percentages of plate modulation a combination of fixed and grid-leak bias can be used, since the variation in rectified grid current is smaller. The grid leak should be by-passed for audio frequencies. The percentage of grid modulation may be regulated by choice of a suitable tap on the modulation-transformer secondary.

The cathode circuit of the modulated stage must be independent of other stages in the transmitter. That is, when directly-heated tubes are modulated their filaments must be supplied from a separate transformer. The filament by-pass condensers should not be larger than about 0.002 μfd., to avoid by-passing the audio-frequency modulation.

Adjustment of Cathode-Modulated Amplifiers

In most respects, the adjustment procedure is similar to that for grid-bias modulation. The critical adjustments are antenna loading, grid bias, and excitation. The proportion of grid-bias to plate modulation will determine the operating conditions.

Adjustments should be made with the aid of an oscilloscope connected in the same way as for grid-bias modulation. With proper antenna loading and excitation, the normal wedge-shaped pattern will be obtained at 100-per-cent modulation. As in the case of grid-bias modulation, too-light antenna loading will cause flattening of the upward-peaks of modulation as also will too-high excitation. The cathode current will be practically constant with or without modulation when the proper operating conditions have been established.

Speech Equipment

In designing speech equipment it is necessary to "work from both ends." That is, we must know, simultaneously, (1) the amount of audio power the modulation system must furnish and (2) the output voltage developed by the microphone when it is spoken into from normal distance (a few inches) with ordinary loudness. It then becomes possible to choose the number and type of amplifier stages needed to generate the required audio power without overloading or distortion anywhere along the line.

The starting point is the microphone.

● MICROPHONES

In this age, no one needs an introduction to the microphone. However, there are several different types of them, considerably different in characteristics. Before considering the various types, it is necessary to define a few terms used in connection with microphones.

The **level** of a microphone is its electrical output for a given sound intensity. Level varies greatly with microphones of different basic types, and also varies between different models of the same type. The output is also greatly dependent on the character of the individual voice (that is, the audio frequencies present in the voice) and the distance of the speaker's lips from the microphone. It decreases approximately as the square of the distance. Because of these variables, only approximate values based on averages of "normal" speaking voices can be given. The values in the following paragraphs are based on close talking; that is, with the microphone about an inch from the speaker's lips.

The **frequency response** or **fidelity** of a microphone is its relative ability to convert sounds of different frequencies into alternating current. With fixed sound intensity at the microphone, the electrical output may vary considerably as the sound frequency is varied. For understandable speech transmission only a limited frequency range is necessary, and intelligible speech can be obtained if the output of the microphone does not vary more than a few decibels at any frequency within a range of about 200 to 2500 cycles. When the variation expressed in terms of decibels is small between two frequency limits, the microphone is said to be **flat** between those limits.

Carbon Microphones

The **carbon microphone** consists of a metal diaphragm placed against an insulating cup containing loosely-packed carbon granules **(microphone button)**. Current from a battery flows through the granules, the diaphragm be-

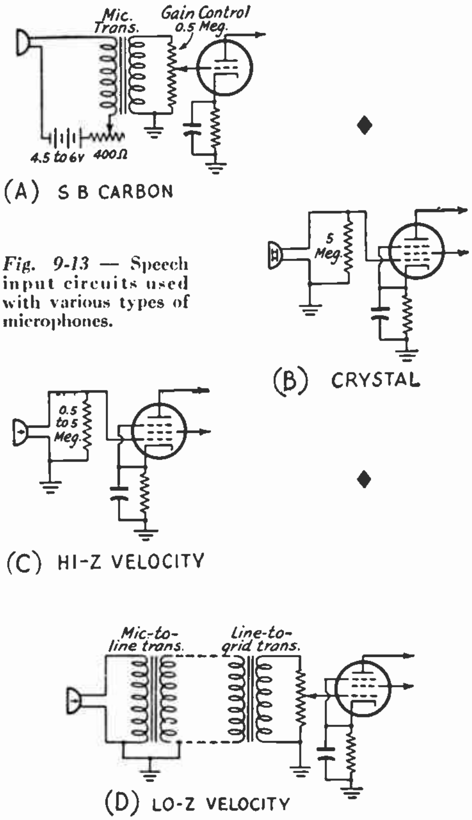


Fig. 9-13 — Speech input circuits used with various types of microphones.

ing one connection and the metal backplate the other. Fig. 9-13A shows connections for carbon microphones. A rheostat is included for adjusting the button current to the value as specified with the microphone. The primary of a transformer is connected in series with the battery and microphone.

As the diaphragm vibrates, its pressure on the granules alternately increases and decreases, causing a corresponding increase and decrease of current flow through the circuit, since the pressure changes the resistance of the mass of granules. The resulting change in the current flowing through the transformer primary causes an alternating voltage, of corresponding frequency and intensity, to be set up in the transformer secondary.

Good-quality carbon microphones give outputs ranging from 0.1 to 0.3 volt across 50 to 100 ohms; that is, across the primary winding of the microphone transformer. With the step-up of the transformer, a peak voltage of between 3 and 10 volts can be assumed to be available at the grid of the amplifier tube. The usual button current is 50 to 100 ma.

Crystal Microphones

The **crystal microphone** makes use of the piezoelectric properties of Rochelle salts crystals. This type of microphone requires no battery or transformer and can be connected directly to the grid of an amplifier tube. It is the most popular type of microphone among amateurs, for these reasons as well as the fact that it has good frequency response and is available in inexpensive models.

The "communications-type" crystal microphone uses a diaphragm mechanically coupled to a crystal. This type of construction gives good sensitivity and adequate frequency response for speech. In higher-fidelity types the sound acts directly on a pair of crystals cemented together, with plated electrodes. The level with the latter construction is considerably less. The input circuit for either model of crystal microphone is shown in Fig. 9-13B.

Although the level of crystal microphones varies with different models, an output of 0.03 volt or so is representative for communication types. The level is affected by the length of the cable connecting the microphone to the first amplifier stage; the above figure is for lengths of 6 or 7 feet. The frequency characteristic is unaffected by the cable, but the load resistance (amplifier grid resistor) does affect it; the lower frequencies are attenuated as the value of load resistance is lowered. A grid-resistor value of at least 1 megohm should be used for reasonably flat response, 5 megohms being a customary figure.

Velocity and Dynamic Microphones

In a **velocity** or **"ribbon" microphone,** the element acted upon by the sound waves is a thin corrugated metallic ribbon suspended between the poles of a magnet. When vibrating, the ribbon cuts the lines of force between the poles, first in one direction and then the other, thus generating an alternating voltage. The movement of the ribbon is proportional to the velocity of the air particles set in motion by the sound.

Velocity microphones are built in two types, high impedance and low impedance, the former being used in most applications. A high-impedance microphone can be directly connected to the grid of an amplifier tube, shunted by a resistance of 0.5 to 5 megohms (Fig. 9-13C). Low-impedance microphones are used when a long connecting cable (75 feet or more) must be employed. In such a case the output of the microphone is coupled to the first amplifier stage through a suitable step-up transformer, as shown in Fig. 9-13D.

The level of the velocity microphone is about 0.03 to 0.05 volt. This figure applies directly to the high-impedance type, and to the low-impedance type when the voltage is measured across the secondary of the coupling transformer.

The **dynamic microphone** somewhat resembles a dynamic loudspeaker. A light-weight voice coil is rigidly attached to a diaphragm, the coil being placed between the poles of a permanent magnet. Sound causes the diaphragm to vibrate, thus moving the coil back and forth between the magnet poles and generating an alternating voltage. The frequency of the generated voltage is proportional to the frequency of the sound waves and the amplitude is proportional to the sound pressure.

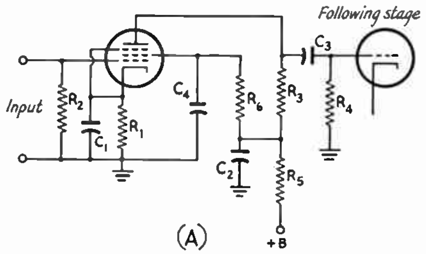
The dynamic microphone usually is built with high-impedance output, suitable for working directly into the grid of an amplifier tube. If the connecting cable must be unusually long, a low-impedance type should be used, with a step-up transformer at the end of the cable.

A small permanent-magnet 'speaker can be used as a dynamic microphone, although the fidelity is not as good as is obtainable with a properly-designed microphone.

THE SPEECH AMPLIFIER

In common terminology, the audio-frequency amplifier stage that actually causes the r.f. carrier output to be varied is called the **modulator,** and all the amplifier stages preceding it comprise the **speech amplifier.** Depending on what sort of modulator is used, the speech amplifier may be called upon to deliver a power output ranging from practically zero (only voltage required) to 20 or 30 watts.

Before starting the design of a speech amplifier, therefore, it is necessary to have selected a



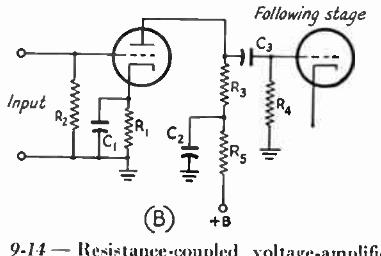


Fig. 9-14 — Resistance-coupled voltage-amplifier circuits. A, pentode; B, triode. Designations are as follows:

C_1 — Cathode by-pass condenser.
C_2 — Plate by-pass condenser.
C_3 — Output coupling condenser (blocking condenser).
C_4 — Screen by-pass condenser.
R_1 — Cathode resistor.
R_2 — Grid resistor.
R_3 — Plate resistor.
R_4 — Next-stage grid resistor.
R_5 — Plate decoupling resistor.
R_6 — Screen resistor.

Values for suitable tubes are given in Table 9-I. Values in the decoupling circuit, C_2R_5, are not critical. R_5 may be about 10% of R_3; an 8- or 10-μfd. electrolytic condenser is usually large enough at C_2.

suitable modulator for the transmitter. This selection must be based on the power required to modulate the transmitter 100 per cent, and this power in turn depends on the type of modulation system selected, as already described. With the modulator picked out, its **driving-power** requirements (audio power required to excite the modulator to full output) can be determined from the tube tables in Chapter Twenty-Five. Generally speaking, it is advisable to choose a tube or tubes for the last stage of the speech amplifier that will be capable of developing at least 50 per cent more power than the rated driving power of the modulator. This will provide a factor of safety so that losses in coupling transformers, etc., will not upset the calculations. A "skimpy" driver, or one designed without a safety factor, usually cannot excite the modulator to full output without being itself overloaded. The inevitable result is speech distortion, generation of unnecessary sidebands, and a "broad" transmitter.

Voltage Amplifiers

If the last stage in the speech amplifier is a Class AB_2 or Class B amplifier, the stage ahead of it must be capable of sufficient power output to drive it. However, if the last stage is a Class AB_1 or Class A amplifier the preceding stage can be simply a voltage amplifier.

From there on back to the microphone, all stages are voltage amplifiers. These are always operated Class A, not only to simplify the design by avoiding driving power, but because just as much *voltage* can be secured from a Class A amplifier as from any other type.

The important characteristics of a voltage amplifier are its **voltage gain,** maximum undistorted **output voltage,** and its **frequency response.** The voltage gain is the voltage-amplification ratio of the stage. The output voltage is the maximum a.f. voltage that can be secured from the stage without distortion; we cannot figure on any greater output voltage than this, no matter what the gain of the stage, without running into the overload region. The amplifier frequency response should be adequate for voice reproduction; this requirement is easily satisfied.

The voltage gain and maximum undistorted output voltage depend on the operating conditions of the amplifier. Data on the popular types of tubes used in speech amplifiers are given in Table 9-I, for resistance-coupled amplification. The output voltage is in terms of *peak* voltage rather than r.m.s.; this method of rating is preferable because it makes the rating independent of the waveform. Exceeding the peak value causes the amplifier to distort, so it is more useful to consider only peak values in working with amplifiers.

Resistance Coupling

Resistance coupling generally is used in voltage-amplifier stages. It is relatively inexpensive, good frequency response can be secured, and there is little danger of hum pick-up from stray magnetic fields associated with heater wiring. It is the only type of coupling suitable for the output circuits of pentodes and high-μ triodes, because with transformers a sufficiently high load impedance cannot be obtained without considerable frequency distortion. Typical circuits are given in Fig. 9-14 and design data in Table 9-I.

Transformer Coupling

Transformer coupling between stages ordinarily is used only when power is to be transferred (in such a case resistance coupling is very inefficient), or when it is necessary to couple between a single-ended and a push-pull stage. Triodes having an amplification factor of 20 or less are used in transformer-coupled voltage amplifiers. With transformer coupling, tubes should be operated under the Class A conditions given in the tube tables in Chapter Twenty-Five.

Representative circuits for coupling single-ended to push-pull stages are shown in Fig. 9-15. The circuit at A combines resistance and transformer coupling, and may be used for exciting the grids of a Class A or AB_1 following stage. The resistance coupling is used to keep the d.c. plate current from flowing through the transformer primary, thereby preventing a reduction in primary inductance below its no-current value; this improves the low-frequency response. With low-μ triodes (6C5, 6J5, etc.), the gain is equal to that with resistance coupling multiplied by the secondary-to-primary turns ratio of the transformer.

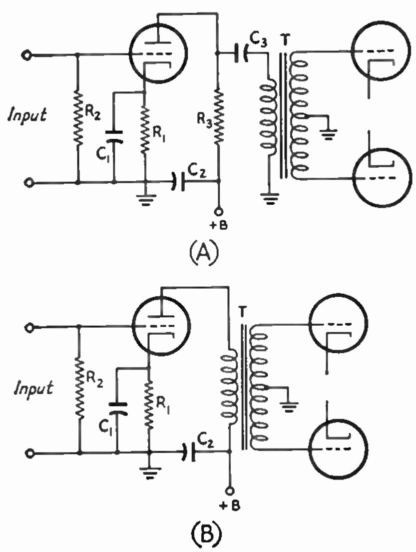


Fig. 9-15 — Transformer-coupled amplifier circuits for driving a push-pull amplifier. A is for resistance-transformer coupling; B for transformer coupling. Designations correspond to those in Fig. 9-14. In A, values can be taken from Table 9-I. In B, the cathode resistor is calculated from the rated plate current and grid bias as given in the tube tables for the particular type of tube used.

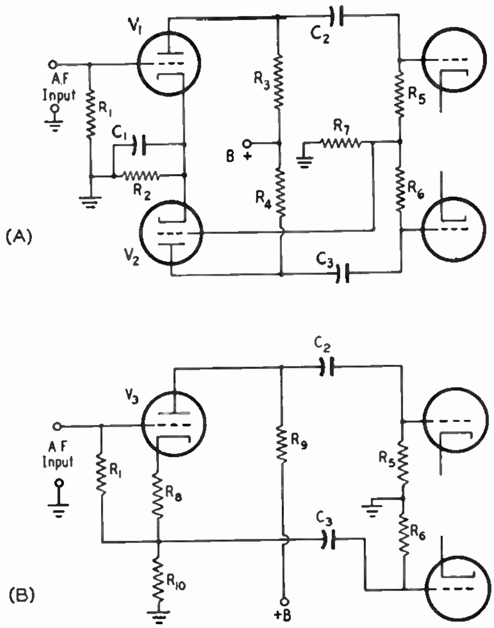


Fig. 9-16 — Self-balancing phase-inverter circuits. V_1 and V_2 may be a double triode such as the 6SN7GT or 6SL7GT. V_3 may be any of the triodes listed in Table 9-I, or one section of a double triode.

R_1 — Grid resistor (1 megohm or less).
R_2 — Cathode resistor; use one-half value given in Table 9-I for tube and operating conditions chosen.
R_3, R_4 — Plate resistor; select from Table 9-I.
R_5, R_6 — Following-stage grid resistor (0.22 to 0.47 megohm).
R_7 — 0.22 megohm.
R_8 — Cathode resistor; select from Table 9-I.
R_9, R_{10} — Each one-half of plate load resistor given in Table 9-I.
C_1 — 10-μfd. electrolytic.
C_2, C_3 — 0.01- to 0.1-μfd. paper.

In B the transformer primary is in series with the plate of the tube, and thus must carry the tube plate current. When the following amplifier operates without grid current, the voltage gain of the stage is practically equal to the μ of the tube multiplied by the transformer ratio. This circuit also is suitable for transferring power (within the capabilities of the tube) to a following Class AB_2 or Class B stage.

Phase Inversion

Push-pull output may be secured with resistance coupling by using "phase-inverter" circuits as shown in Fig. 9-16.

The circuit shown in Fig. 9-16A is known as the "self-balancing" type. The amplified voltage from V_1 appears across R_5 and R_7 in series. The drop across R_7 is applied to the grid of V_2, and the amplified voltage from V_2 appears across R_6 and R_7 in series. This voltage is 180 degrees out of phase with the voltage from V_1,

TABLE 9-I — RESISTANCE-COUPLED VOLTAGE-AMPLIFIER DATA

Data are given for a plate supply of 300 volts. Departures of as much as 50 per cent from this supply voltage will not materially change the operating conditions or the voltage gain, but the output voltage will be in proportion to the new voltage. Voltage gain is measured at 400 cycles; condenser values given are based on 100-cycle cut-off. For increased low-frequency response, all condensers may be made larger than specified (cut-off frequency in inverse proportion to condenser values provided all are changed in the same proportion). A variation of 10 per cent in the values given has negligible effect on the performance.

	Plate Resistor Megohms	Next-Stage Grid Resistor Megohms	Screen Resistor Megohms	Cathode Resistor Ohms	Screen By-pass μfd.	Cathode By-pass μfd.	Blocking Condenser μfd.	Output Volts (Peak)[1]	Voltage Gain[2]
6SJ7	0.1	0.1	0.35	500	0.10	11.6	0.019	72	67
		0.25	0.37	530	0.09	10.9	0.016	96	98
		0.5	0.47	590	0.09	9.9	0.007	101	104
	0.25	0.25	0.89	850	0.07	8.5	0.011	79	139
		0.5	1.10	860	0.06	7.4	0.004	88	167
		1.0	1.18	910	0.06	6.9	0.003	98	185
	0.5	0.5	2.0	1300	0.06	6.0	0.004	64	200
		1.0	2.2	1410	0.05	5.8	0.002	79	238
		2.0	2.5	1530	0.04	5.2	0.0015	89	263
6J7, 7C7	0.1	0.1	0.44	500	0.07	8.5	0.02	55	61
		0.25	0.5	450	0.07	8.3	0.01	81	82
		0.5	0.53	600	0.06	8.0	0.006	96	94
	0.25	0.25	1.18	1100	0.04	5.5	0.008	81	104
		0.5	1.18	1200	0.04	5.4	0.005	104	140
		1.0	1.45	1300	0.05	5.8	0.005	110	185
	0.5	0.5	2.45	1700	0.04	4.2	0.005	75	161
		1.0	2.9	2200	0.04	4.1	0.003	97	200
		2.0	2.95	2300	0.04	4.0	0.0025	100	230
6AU6, 6SH7	0.1	0.1	0.2	500	0.13	18.0	0.019	76	109
		0.22	0.24	600	0.11	16.4	0.011	103	145
		0.47	0.26	700	0.11	15.3	0.006	129	168
	0.22	0.22	0.42	1000	0.1	12.4	0.009	92	164
		0.47	0.5	1000	0.098	12.0	0.007	108	230
		1.0	0.55	1100	0.09	11.0	0.003	122	262
	0.47	0.47	1.0	1800	0.075	8.0	0.0045	94	248
		1.0	1.1	1900	0.065	7.6	0.0028	105	318
		2.2	1.2	2100	0.06	7.3	0.0018	122	371
6AQ6, 6AT6, 6Q7, 6SL7GT (one triode)	0.1	0.1	—	1500	—	4.4	0.027	40	34
		0.22	—	1800	—	3.6	0.014	54	38
		0.47	—	2100	—	3.0	0.0065	63	41
	0.22	0.22	—	2600	—	2.5	0.013	51	42
		0.47	—	3200	—	1.9	0.0065	65	46
		1.0	—	3700	—	1.6	0.0035	77	48
	0.47	0.47	—	5200	—	1.2	0.006	61	48
		1.0	—	6300	—	1.0	0.0035	74	50
		2.2	—	7200	—	0.9	0.002	85	51
6F5, 6SF5, 7B4	0.1	0.1	—	1300	—	5.0	0.025	33	42
		0.25	—	1600	—	3.7	0.01	43	49
		0.5	—	1700	—	3.2	0.006	48	52
	0.25	0.25	—	2600	—	2.5	0.01	41	56
		0.5	—	3200	—	2.1	0.007	54	63
		1.0	—	3500	—	2.0	0.004	63	67
	0.5	0.5	—	4500	—	1.5	0.006	50	65
		1.0	—	5400	—	1.2	0.004	62	70
		2.0	—	6100	—	0.93	0.002	70	70
6SC7[3] (one triode)	0.1	0.1	—	750	—	—	0.033	35	29
		0.25	—	930	—	—	0.014	50	34
		0.5	—	1040	—	—	0.007	54	36
	0.25	0.25	—	1400	—	—	0.012	45	39
		0.5	—	1680	—	—	0.006	55	42
		1.0	—	1840	—	—	0.003	64	45
	0.5	0.5	—	2330	—	—	0.006	50	45
		1.0	—	2980	—	—	0.003	62	48
		2.0	—	3280	—	—	0.002	72	49
6J5, 7A4, 7N7, 6SN7GT (one triode)	0.05	0.05	—	1020	—	3.56	0.06	41	13
		0.1	—	1270	—	2.96	0.034	51	14
		0.25	—	1500	—	2.15	0.012	60	14
	0.1	0.1	—	1900	—	2.31	0.035	43	14
		0.25	—	2440	—	1.42	0.0125	56	14
		0.5	—	2700	—	1.2	0.0065	64	14
	0.25	0.25	—	4590	—	0.87	0.013	46	14
		0.5	—	5770	—	0.64	0.0075	57	14
		1.0	—	6950	—	0.54	0.004	64	14
6C4	0.047	0.047	—	870	—	4.1	0.065	38	12
		0.1	—	1200	—	3.0	0.034	52	12
		0.22	—	1500	—	2.4	0.016	68	12
	0.1	0.1	—	1900	—	1.9	0.032	44	12
		0.22	—	3000	—	1.3	0.016	68	12
		0.47	—	4000	—	1.1	0.007	80	12
	0.22	0.22	—	5300	—	0.9	0.015	57	12
		0.47	—	800	—	0.52	0.007	82	12
		1.0	—	11000	—	0.46	0.0035	92	12

[1] Voltage across next-stage grid resistor at grid-current point.
[2] At 5 volts r.m.s. output.
[3] Cathode-resistor values are for phase-inverter service.

thus giving push-pull output. The part that appears across R_7 therefore opposes the voltage from V_1 across R_7, thus reducing the signal applied to the grid of V_2. The negative feed-back so obtained tends to regulate automatically the voltage applied to the phase-inverter tube so that the output voltages from both tubes are substantially equal — as they must be for distortionless reproduction. The self-balancing circuit also has the advantage of compensating for variations in the characteristics of the two tubes. The gain is slightly less than twice the gain of a single-tube amplifier using the same operating conditions.

The single-tube circuit shown in Fig. 9-16B also is inherently balanced. In this case the plate load resistor is divided into two equal parts, R_9 and R_{10}, one being connected to the plate in the normal way and the other between cathode and ground. Since the voltages at the plate and cathode are 180 degrees out of phase, the grids of the following tubes are fed equal a.f. voltages in push-pull. The grid return of V_3 is made to the junction of R_8 and R_{10} so normal bias will be applied to the grid. This circuit is highly degenerative because of the way R_{10} is connected. The voltage gain is less than 2 even when a high-μ triode is used at V_3.

Gain Control

A means for varying the over-all gain of the amplifier is a practical necessity. Without it, there would be no way to keep the final output down to the proper level for modulating the transmitter except to talk at just the right intensity. The common method of gain control is to adjust the value of a.c. voltage applied to the grid of one of the amplifiers by means of a voltage divider or potentiometer.

The gain-control potentiometer should be near the input end of the amplifier, at a point where the a.c. voltage level is so low that there is no danger of overloading in the stages ahead of the gain control. With carbon microphones the gain control may be placed directly across the microphone-transformer secondary. With other types of microphones, however, the gain control usually will affect the frequency response of the microphone when connected directly across it. The control therefore is usually placed in the grid circuit of the second stage.

DESIGNING THE SPEECH AMPLIFIER

The steps in designing a speech amplifier are as follows:

1) Determine the power needed to modulate the transmitter and select the modulator. In the case of plate modulation, this will nearly always be a Class B amplifier. Select a suitable tube type and determine from the tube tables in Chapter Twenty-Five the driving power required.

2) As a safety factor, multiply the required driver power by at least 1.5.

3) Select a tube, or pair of tubes, that will deliver the power determined in the second step. This is the last speech-amplifier stage. Receiver-type power tubes can be used (beam tubes such as the 6L6 may be needed in some cases) so the receiving-tube tables in Chapter Twenty-Five may be consulted. If the speech amplifier is to drive a Class B modulator, use a Class A or AB_1 amplifier if it will give enough power output.

4) If the last speech-amplifier stage *has* to operate Class AB_2, use a medium-μ triode (such as the 6J5 or corresponding types) to drive it. In the extreme case of driving 6L6s to maximum output, two triodes should be used in push-pull in the driver. In either case transformer coupling will have to be used, and transformer manufacturers' catalogs should be consulted for a suitable type.

5) If the last speech-amplifier stage operates Class A or AB_1, it may be driven by a voltage amplifier. If the last stage is push-pull, the driver may be a single tube coupled through a transformer with a balanced secondary, or may be a dual-triode phase inverter. Determine the signal voltage required for full output from the last stage. If the last stage is a single-tube Class A amplifier, the peak signal is equal to the grid-bias voltage; if push-pull Class A, the peak signal voltage is equal to twice the grid bias; if Class AB_1, twice the bias voltage when fixed bias is used; if cathode bias is used, twice the bias figured from the cathode resistance and the no-signal plate current.

6) From Table 9-I, select a tube capable of giving the required output voltage and note its rated voltage gain. A double-triode phase inverter (Fig. 9-16A) will have approximately twice the output voltage and twice the gain of one triode operating as an ordinary amplifier. If the driver is to be transformer-coupled to the last stage, select a medium-μ triode and calculate the gain and output voltage as previously described.

7) Divide the voltage required to drive the last stage by the gain of the preceding stage. This gives the peak voltage required at the grid of the next-to-the-last stage.

8) Find the output voltage, under ordinary conditions, of the microphone to be used. This information should be obtained from the manufacturer's catalog. If not available, the figures given in the section on microphones in this chapter will serve.

9) Divide the voltage found in (7) by the output voltage of the microphone. The result is the over-all gain required from the microphone to the grid of the next-to-the-last stage. To be on the safe side, double or triple this figure.

10) From Table 9-I, select a combination of tubes whose gains, when multiplied together, give approximately the figure arrived at in (9). These amplifiers will be used in cascade. In

general, if high gain is required it is advisable to use a pentode for the first speech-amplifier stage, but it is *not* advisable to use a second pentode because of the possibility of feed-back and self-oscillation. In most cases a triode will give enough gain, as a second stage, to make up the total gain required. If not, a third stage, also a triode, may be used.

SPEECH-AMPLIFIER CONSTRUCTION

Once a suitable circuit has been selected for a speech amplifier, the construction problem resolves itself into avoiding two difficulties — excessive hum, and unwanted feed-back. For reasonably humless operation, the hum voltage should not exceed about 1 per cent of the maximum audio output voltage — that is, the hum should be about 40 db. below the output level. Unwanted feed-back, if negative, will reduce the gain below the calculated value; if positive, is likely to cause self-oscillation or "howls." Feed-back can be minimized by isolating each stage with "decoupling" resistors and condensers, by avoiding layouts that bring the first and last stages near each other, and by shielding of "hot" points in the circuit, such as grid leads in low-level stages.

Speech-amplifier equipment, especially voltage amplifiers, should be constructed on metal chassis, with all wiring kept below the chassis to take advantage of the shielding afforded. Exposed leads, particularly to the grids of low-level high-gain tubes, are likely to pick up hum from the electrostatic field that usually exists in the vicinity of house wiring. Even with the chassis, additional shielding of the input circuit of the first tube in a high-gain amplifier usually is necessary. In addition, such circuits should be separated as much as possible from power-supply transformers and chokes and also from any audio transformers that operate at fairly-high power levels; this will minimize magnetic coupling to the grid circuit and thus reduce hum or audio-frequency feed-back. It is always a safe plan, although not an absolutely necessary one, to build the speech amplifier and its power supply as separate units.

If a low-level microphone such as the crystal type is used, the microphone, its connecting cable, and the plug or connector by which it is attached to the speech amplifier, all should be shielded. The microphone and cable usually are constructed with suitable shielding. The cable shield should be connected to the speech-amplifier chassis, and it is advisable — as well as usually necessary — to connect the chassis to a ground such as a water pipe.

Heater wiring should be kept as far as possible from grid leads, and either the center-tap or one side of the heater-transformer secondary winding should be connected to the chassis. If the center-tap is grounded, the heater leads to each tube should be twisted together to reduce the magnetic field from the heater current. With either type of connection, it is advisable to lay heater leads in the corner formed by a fold in the chassis, bringing them out from the corner to the tube socket by the shortest possible path.

In a high-gain amplifier it is sometimes helpful if the first tube has its grid connection brought out to a top cap rather than to a base pin; in the latter type the grid lead is exposed to the heater leads inside the tube and hence may pick up more hum. With the top-cap tubes, complete shielding of the grid lead and grid cap is a necessity.

When metal tubes are used, always ground the shell connection to the chassis. Glass tubes used in the low-level stages of high-gain amplifiers must be shielded; tube shields are obtainable for that purpose. It is a good plan to enclose the entire amplifier in a metal box, or at least provide it with a cane-metal cover, to avoid feed-back difficulties caused by the r.f. field of the transmitter; r.f. picked up on exposed wiring leads or tube elements causes overloading, distortion, and frequently oscillation.

When using paper condensers as by-passes, be sure that the terminal marked "outside foil" is connected to ground. This utilizes the outside foil of the condenser as a shield around the "hot" foil. When paper condensers are used as coupling condensers between stages, always connect the outside-foil terminal to the side of the circuit having the lowest impedance to ground. Usually, this will be the plate side rather than the following-grid side.

INCREASING THE EFFECTIVENESS OF THE 'PHONE TRANSMITTER

The design principles outlined so far in this section are perfectly straightforward and apply to amplifiers designed for any purpose. However, the effectiveness of an amateur 'phone transmitter can be increased to a remarkable extent by taking advantage of speech characteristics and of the requirements in *voice* communication.

Measures that may be taken to make the modulation more effective include band compression (filtering), volume compression, and speech clipping.

Compressing the Frequency Band

Most of the intelligibility in speech is contained in the medium band of frequencies; that is, between about 500 and 2500 cycles. On the other hand, the major portion of speech power is normally concentrated below 500 cycles. It is these low frequencies that modulate the transmitter most heavily. If they are eliminated, the frequencies that carry most of the actual communication can be increased in amplitude without exceeding 100-per-cent modulation, and the effectiveness of the transmitter is correspondingly increased.

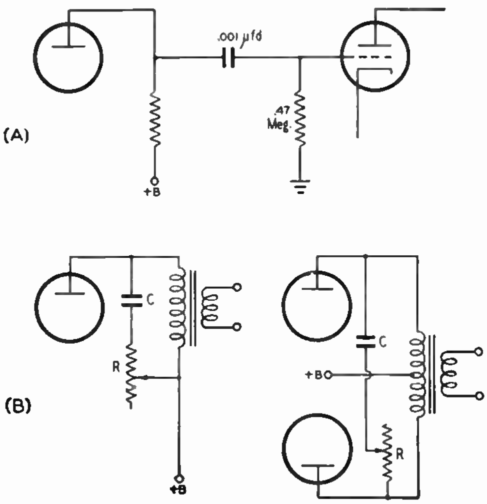


Fig. 9-17 — A, use of a small coupling condenser to reduce low-frequency response; B, tone-control circuits for reducing high-frequency response. Values for C and R are discussed in the text; 0.01 μfd. and 25,000 ohms are typical.

One simple way to reduce low-frequency response is to use small values of coupling capacitance between resistance-coupled stages, as shown in Fig. 9-17A. A time constant of 0.0005 second for the coupling condenser and following-stage grid resistor will have little effect on the amplification at 500 cycles, but will practically halve it at 100 cycles. In two cascaded stages the gain will be down about 5 db. at 200 cycles and 10 db. at 100 cycles. When the grid resistor is ½ megohm a coupling condenser of 0.001 μfd. will give the required time constant.

The high-frequency response can be reduced by using "tone control" methods, utilizing a condenser in series with a variable resistor connected across an audio impedance at some point in the speech amplifier. The best spot for the tone control is across the primary of the output transformer of the speech amplifier, as in Fig. 9-17B. The condenser should have a reactance at 1000 cycles about equal to the load resistance required by the amplifier tube or tubes, while the variable resistor in series may have a value equal to four or five times the load resistance. The control can be adjusted while listening to the amplifier, the object being to cut the high-frequency response as much as possible without unduly sacrificing intelligibility.

Compressing the audio-frequency band not only puts more modulation power in the optimum frequency band but also reduces hum, because the low-frequency response is reduced, and helps reduce the width of the channel occupied by the transmission, because of the reduction in the amplitude of the high audio frequencies.

Volume Compression

It is obviously desirable to modulate the transmitter as completely as possible — without, of course, overmodulating and setting up spurious sidebands. However, it is difficult to maintain constant voice intensity when speaking into the microphone. To overcome this variable output level, it is possible to use automatic gain control that follows the *average* (not instantaneous) variations in speech amplitude. This can be done by rectifying and filtering some of the audio output and applying the rectified and filtered d.c. to a control electrode in an early stage in the amplifier.

A practical circuit for this purpose is shown in Fig. 9-18. The rectifier must be connected, through the transformer, to a tube capable of delivering some power output (a small part of the output of the power stage may be used) or else a separate power amplifier for the rectifier circuit alone may have its grid connected in parallel with that of the last voltage amplifier.

Resistor R_4, in series with R_5 across the plate supply, provides an adjustable positive bias on the rectifier cathodes. This prevents the limiting action from beginning until a desired microphone input level is reached. R_2, R_3, C_2, C_3 and C_4 filter the audio frequencies from the rectified output. The output of the rectifier may be connected to the suppressor grid of a pentode first stage of the speech amplifier.

A step-down transformer with a turns ratio such as to give about 50 volts when its primary is connected to the output circuit of the power stage should be used. If a transformer having a center-tapped secondary is not available, a half-wave rectifier may be used instead of the full-wave circuit shown, but it will be harder to get satisfactory filtering.

The over-all gain of the system must be high enough so that full output can be secured at a moderately low voice level.

Speech Clipping and Filtering

Earlier in this chapter it was pointed out that with sine-wave 100-per-cent modulation the average power increases to 150 per cent of the unmodulated carrier power, but that in speech waveforms the average power content is considerably less than in a sine wave, when

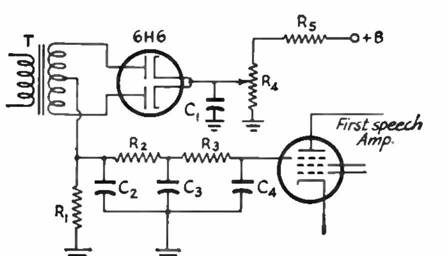


Fig. 9-18 — Speech-amplifier output-limiting circuit. C_1, C_2, C_3, C_4 — 0.1-μfd.; R_1, R_2, R_3 — 0.22 megohm; R_4 — 25,000-ohm pot.; R_5 — 0.1 megohm; T — see text.

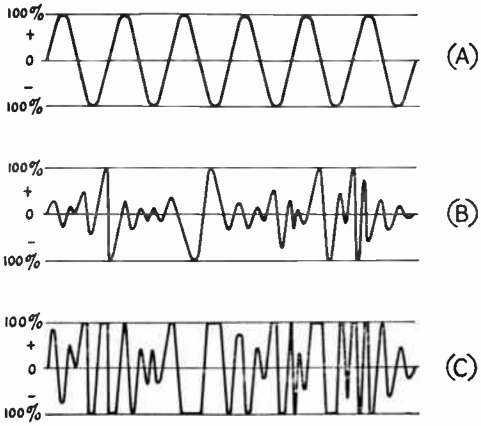


Fig. 9-19 — The normal speech wave (B) has high peaks but low average energy content. When the peaks are clipped the signal may be increased to a considerably-higher power level without causing overmodulation (C).

both waveforms have the same *peak* amplitude. Nevertheless, it is the peak conditions that count in modulation. This is shown in the drawings of Fig. 9-19. The upper drawing, A, represents a sine wave having a maximum amplitude that just modulates a given transmitter 100 per cent. The same maximum amplitude will modulate the same transmitter 100 per cent regardless of the waveform of the modulating signal. The speech wave at B, therefore, also represents 100-per-cent modulation.

Now suppose that the amplitude of the wave shown at B is increased so that its power is comparable with — or even higher than — the power in a sine wave, but that everything above the 100-per-cent modulation mark is cut off. We then have a wave such as is shown at C, which is the wave at B increased in amplitude but with its peaks "clipped." This signal will not modulate the transmitter more than 100 per cent, but the voice power will be several times as great. The wave is not exactly like the one at B, so the result will not sound exactly like the original. However, such clipping can be used to secure a worth-while increase in modulation power without sacrificing *intelligibility*. The clipping can be done in the speech amplifier, and once the system is properly adjusted *it will be impossible to overmodulate the transmitter* no matter how much gain is used ahead of the clipper — because the clipper will hold the maximum output amplitude to the same value no matter what the amplitude of the signal applied to it.

But by itself the clipper is not enough. Although the clipping takes place in the audio system, the signal applied to the modulated r.f. amplifier has practically the same waveshape that the modulation envelope *would have had* if the signal were unclipped and the transmitter were badly overmodulated. In other words, clipping generates the same high-order harmonics that overmodulation does. It is therefore necessary to prevent the higher audio frequencies from reaching the modulator. In other words, the frequencies above those needed for intelligible speech must be filtered out, *after* clipping and *before* modulation. The filter required for this purpose should have relatively little attenuation at frequencies below about 2500 cycles, but very great attenuation for all frequencies above 3000 cycles.

It is possible to use as much as 25 db. of clipping before intelligibility is lost; that is, if the original peak amplitude is 10 volts, the signal can be clipped to such an extent that the resulting maximum amplitude is less than one volt. If the original 10-volt signal represented the amplitude that caused 100-per-cent modulation on peaks, the clipped and filtered signal can then be amplified up to the same 10-volt peak level for modulating the transmitter, with a very considerable increase in modulation power.

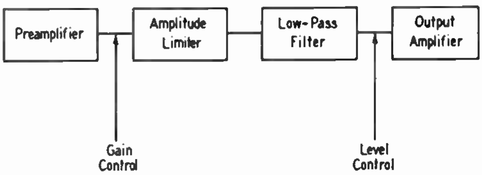


Fig. 9-20 — Block diagram of speech-clipping and filtering amplifier.

There is a loss in naturalness with "deep" clipping, even though the voice is highly intelligible. With moderate clipping levels (6 db. or so) there is almost no perceptible change in "quality" but the voice power is four or five times as great as in ordinary modulation.

Before drastic clipping can be used, the speech signal must be amplified up to 10 times more than is necessary for normal modulation. Also, the hum and noise must be much lower than the tolerable level in ordinary amplification, because the noise in the output of the amplifier increases in proportion to the gain.

The clipper-filter system is shown in block form in Fig. 9-20. The limiter is a peak-limiting rectifier of the same general type that is used in receiver noise limiters. It must clip both positive and negative peaks. The gain control sets the amplitude at which clipping starts. Following the low-pass filter for eliminating the harmonic distortion frequencies is a second gain control, the "level" control. This control is set initially so that the amplitude-limited output of the clipper-filter modulates the transmitter 100 per cent. Thereafter it need not be touched. The clipper-filter system is consequently an automatic "overmodulation-preventer," and is a worth-while addition to the transmitter on that account even though deep clipping is seldom used.

Practical circuits are illustrated in a speech amplifier described later in this chapter.

Speech Amplifier with Push-Pull Triode Output

The speech amplifier shown in Fig. 9-21 is a general-purpose unit of straightforward design. Using a pair of power triodes in the output stage, it is capable of an actual undistorted output power of about 8 watts. It can therefore be used to drive a Class B modulator of moderate power output. It is also suitable for use as a grid-bias modulator for high-power transmitters. The gain of the amplifier is ample for the ordinary communications-type crystal microphone.

As shown in the circuit diagram, Fig. 9-22, the amplifier has a pentode first stage using a 6SJ7. A medium-μ triode, a 6J5, is used in the second stage. The gain control is in the grid circuit of this tube. The third stage uses a 6SL7GT in the self-balancing phase-inverter circuit, to obtain push-pull output for the grids of the output tubes. The final stage has two 6B4Gs in push-pull, operated Class AB_1. The power supply for the amplifier is included on the same chassis.

The circuits of individual stages are basically as described earlier in this chapter. R_6 and R_{10} are decoupling resistors in the 6SJ7 and 6J5 stages, respectively, to prevent unwanted feed-back. These resistors, in combination with C_3 and C_7, also provide some additional power-supply hum filtering for the first two stages where the signal level is low. Condenser C_5, which is shunted across the gain control, R_5, when S_1 is closed, serves to reduce the gain at frequencies above about 2500 cycles. This, as explained earlier in this chapter, is desirable because it reduces the width of the channel occupied by the transmitter. R_{17} and C_{12} are the cathode-bias resistor and by-pass condenser, respectively, for the output stage. C_{12} should not be omitted unless the output stage operating conditions are changed so that the amplifier operates purely Class A. When the plate current varies, as it does in Class AB_1 operation, the varying current through R_{17} will introduce considerable distortion unless the resistor is by-passed by a low-impedance condenser.

In the power-supply circuit, S_3 is used for shutting off the plate voltage while leaving the heater power on the tubes. A two-section condenser-input smoothing filter is used. The plate voltage for the output amplifier is taken from the first section; this makes the voltage available for the plates somewhat higher because it avoids the voltage drop through L_2. Hum at this point is inconsequential because of the high power level.

Parts Layout

The speech section occupies the left-hand side of the chassis and the power-supply section the right. Controls along the front chassis edge are the tone-control switch, S_1, gain control, R_5, microphone connector, "B" switch, S_3, and a.c. switch, S_2. The 6SJ7 is behind the microphone connector on the chassis, and the 6J5 is to its left, near the gain control. The 6SL7 phase inverter and 6B4G output tubes are located behind the 6J5.

At the right, the power transformer is at the rear of the chassis, the 5Y3GT rectifier in front, and the first filter choke, L_1, is to the left of the rectifier tube. The output transformer is at the rear center of the chassis.

The bottom view shows the cathode resistor, R_{17}, for the 6B4Gs at the lower right, together with its by-pass condenser, C_{12}. Just above is the second filter choke, L_2. The filter condensers, C_{13}, C_{14} and C_{15}, are the larger tubular units located to the left. The resistors and condensers associated with individual stages are grouped about the appropriate tube sockets. The terminals of the output transformer, T_1, project through a cut-out in the chassis, and secondary leads are brought out to a terminal strip.

A shielded lead should be used from the microphone connector to

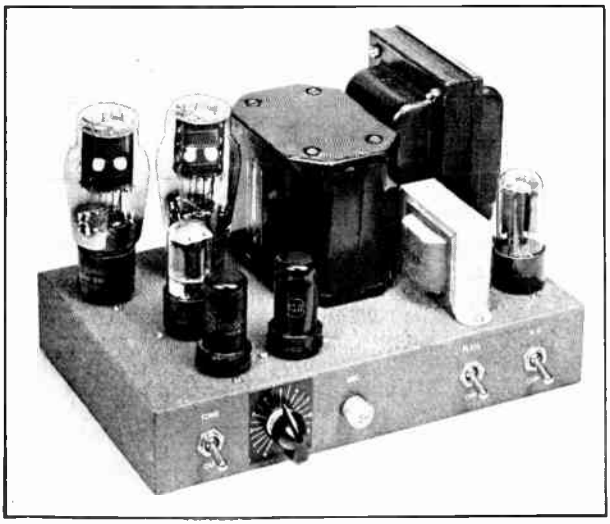


Fig. 9-21 — This amplifier uses 6B4Gs (equivalent to 6A3s) as output tubes and will deliver 8 watts of undistorted power. It is complete with power supply on a 7 × 11 × 2-inch chassis.

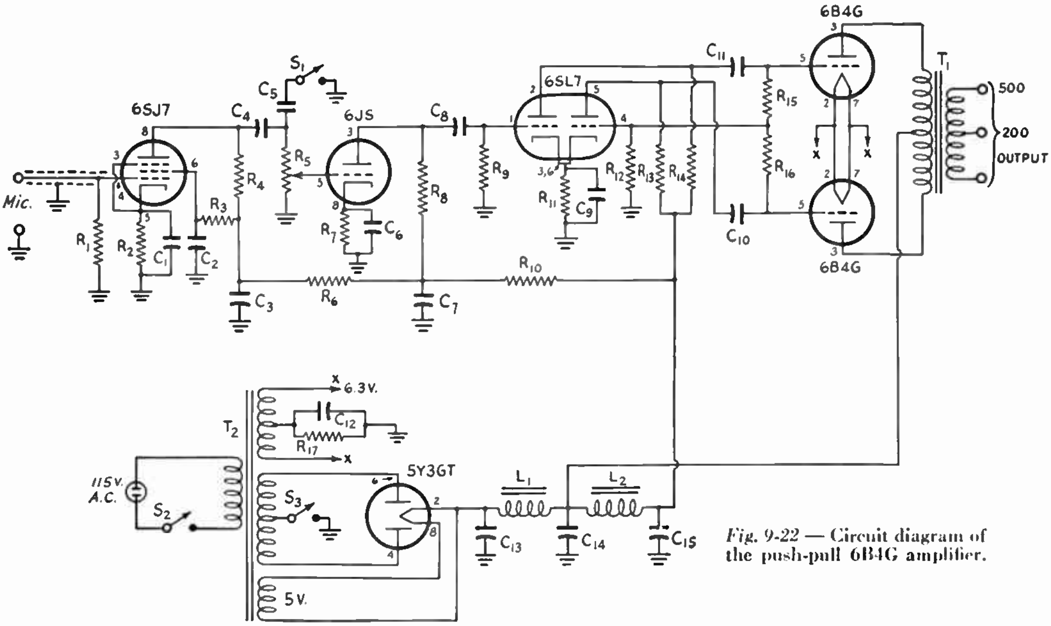


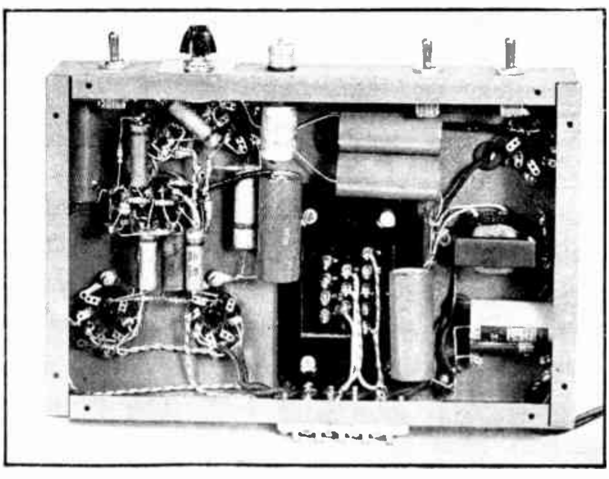
Fig. 9-22 — Circuit diagram of the push-pull 6B4G amplifier.

C_1, C_6, C_9 — 20-μfd. 25-volt electrolytic.
C_2 — 0.1-μfd. 400-volt paper.
C_3, C_7, C_{13}, C_{14}, C_{15} — 10-μfd. 450-volt electrolytic.
C_4, C_8, C_{10}, C_{11} — 0.01-μfd. 600-volt paper.
C_5 — 0.001-μfd. 500-volt mica.
C_{12} — 50-μfd. 100-volt electrolytic.
R_1 — 1 megohm, ½ watt.
R_2, R_7 — 1500 ohms, ½ watt.
R_3 — 1.5 megohms, ½ watt.
R_4, R_{12}, R_{13}, R_{14}, R_{15}, R_{16} — 0.22 megohm, ½ watt.
R_5 — 0.5-megohm volume control.
R_6 — 47,000 ohms, ½ watt.
R_8 — 82,000 ohms, ½ watt.
R_9 — 0.47 megohm, ½ watt.
R_{10} — 10,000 ohms, 1 watt.
R_{11} — 1500 ohms, 1 watt.
R_{17} — 750 ohms, 10 watts.
L_1 — 8-hy. 160-ma. filter choke (UTC R-20).
L_2 — 10-hy. 35-ma. filter choke (UTC R-55).
S_1, S_2, S_3 — S.p.s.t. toggle.
T_1 — Output transformer, p.p. plates (5000 ohms) to line (UTC PA-16).
T_2 — 700 volts c.t., 110 ma.; 5 volts, 3 amp.; 6.3 volts, 4.5 amp. (Stancor P-4080).

the grid prong on the 6SJ7 socket, but there are otherwise no special constructional precautions to observe — other than those mentioned in the section on general considerations in speech-amplifier construction.

The output transformer shown in the photographs is designed for working into a 500- or 200-ohm line. This type of transformer may be used when the speech amplifier is located at some distance from the Class B modulator or other unit it is to drive. If desired, a Class B input transformer can be substituted at T_1. In that case, the leads to the modulator-tube grids should be shielded as a precaution against hum or r.f. pick-up. The transformer selected should be designed for working from a 5000-ohm plate-to-plate load to the grids of the modulator tubes selected.

Fig. 9-23 — Bottom view of the push-pull 6B4G amplifier. Output-transformer terminals are brought out to a connection strip on the rear edge of the chassis.



A Clipper-Filter Speech Amplifier

The amplifier shown in Fig. 9-24 has a usable output of about 4 watts (sine wave) and includes a clipper-filter for increasing the effectiveness of the modulator and for confining the channel-width to the frequencies needed for intelligible speech. The output stage uses a 6V6 with negative feed-back: this reduces the effective plate resistance of the tube to a low value. The unit therefore can be used to drive a Class B modulator that does not require more than 4 watts on the grids. It can also be used as a complete modulator unit for grid-bias modulation.

As shown in the circuit diagram, Fig. 9-25, the first tube is a 6SJ7. The second stage is one section of a 6SL7GT. With S_3 thrown to the right-hand position, the output of this stage is connected to the grid of a 6J5, which in turn drives the 6V6. Under these conditions the amplifier operates conventionally and has fairly wide frequency response. With S_3 thrown to the left, the output of the first 6SL7GT section is fed to the 6AL5 clipper, and the clipped output is then fed to the grid of the second section of the 6SL7GT. The output of this tube goes through a low-pass filter and thence through a second gain control, R_{15}, to the grid of the 6J5. Thus the clipper-filter feature can be used or not as desired.

The first two stages are resistance-coupled amplifiers following ordinary practice. In the last stage, use is made of the center-tap on the primary of the output transformer to obtain feed-back voltage that is applied to the grid of the 6V6 through the plate resistor, R_{13}, of the 6J5. If a different type of transformer is used, not having a center-tap, a voltage divider can be connected across the primary to obtain the feed-back voltage, as described in the section on negative feed-back in this chapter.

The amplifier has its own power supply, as shown in the diagram and photographs.

Circuit Notes

The clipper circuit uses two diodes, one to clip positive and the other to clip negative peaks, in shunt with a load resistor, R_{11}. The diodes are biased so that they are nonconducting until the signal amplitude reaches about 2 volts. When conducting, the diode resistance is low compared to the resistance of R_{11}, and also compared to the series resistance R_{10}. Under these conditions, all of the voltage in excess of the 2-volt bias appears as a voltage drop in R_{10} (and in the plate resistance of the preceding stage), with the result that the voltage across R_{11} cannot exceed 2 volts.

For convenience, the bias for the diodes is taken from the cathode resistor of the 6V6 by a voltage-dividing arrangement. As shown in Fig. 9-25, the plate of one diode is connected to ground, R_{11} is returned to a point 2 volts above ground, and the cathode of the second diode is returned to a point 4 volts above ground. This makes the plate of each diode 2 volts negative with respect to its own cathode.

The filter shown in Fig. 9-26 is constructed of standard components, the chokes being 125-mh. units usually sold as r.f. chokes. The design of a filter using this value of inductance requires a fairly high capacitance and a low value of load resistance. The constants listed give a sharp cut-off between 2500 and 3000 cycles, with very large attenuation (averaging 45 db. below the response at 1000 cycles) at all frequencies above 3000 cycles. However, the low value of load or terminating resistor, 2000 ohms, greatly decreases the voltage amplification of the 6SL7GT section as compared to what could be obtained with a normal load. The over-all gain with R_{15} at maximum is about the same as with S_3 in the "normal-amplifier" position, despite the extra stage, when the input signal is below the clipping level. Once clipping begins, of course, the output voltage cannot rise above the clipping level no matter how high the amplitude of the input signal.

Construction

The amplifier is built on a 6 × 14 × 3-inch chassis. The input end of the speech amplifier is at the left end and the power supply is at the right. A shield is placed over the 6SL7GT to prevent hum pick-up and to protect the

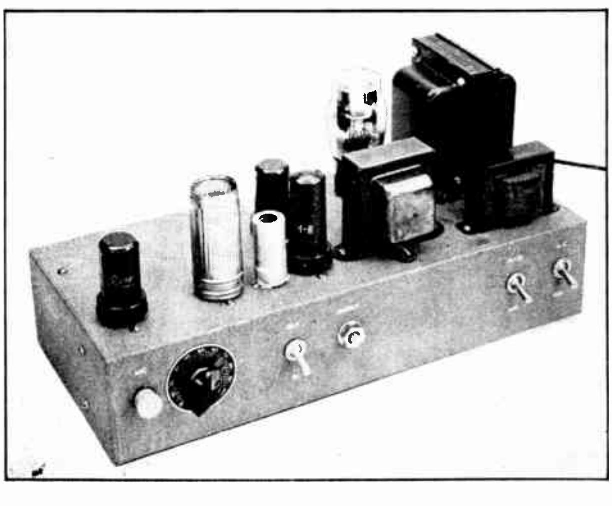


Fig. 9-24 — **A 4-watt output amplifier with speech clipping and filtering. It uses a 6V6 output tube with negative feed-back, and has its power supply on the same chassis.**

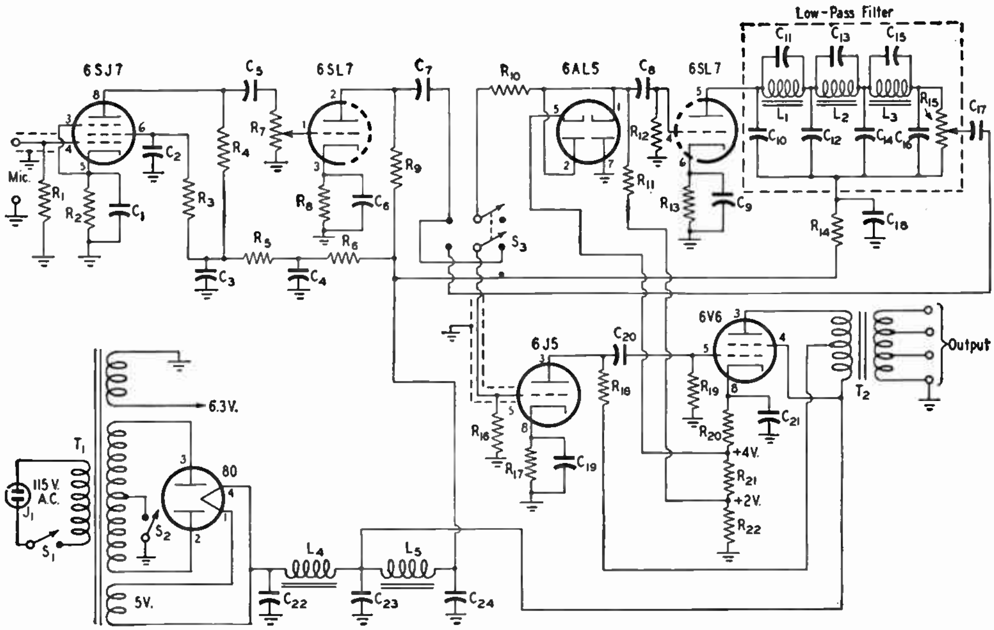


Fig. 9-25 — Circuit diagram of the clipper-filter speech amplifier.

C_1, C_6, C_9, C_{19} — 10-μfd. 25-volt electrolytic.
C_2 — 0.1-μfd. 400-volt paper.
C_3, C_4, C_{18}, C_{22}, C_{23} — 8-μfd. 450-volt electrolytic.
C_5, C_7, C_8, C_{17}, C_{20} — 0.01-μfd. 600-volt paper.
C_{10}, C_{11}, C_{13} — 0.015-μfd. paper.
C_{12} — 0.03-μfd. paper.
C_{14} — 0.05-μfd. paper.
C_{15} — 0.003-μfd. mica.
C_{16} — 0.06-μfd. paper.
C_{21} — 50-μfd. 50-volt electrolytic.
C_{24} — 16-μfd. 450-volt electrolytic.
R_1 — 1 megohm, ½ watt.
R_2, R_{13} — 1000 ohms, ½ watt.
R_3 — 1.2 megohms, ½ watt.
R_4 — 0.22 megohm, ½ watt.
R_5, R_{10} — 47,000 ohms, ½ watt.
R_6 — 0.1 megohm, ½ watt.
R_7 — 2-megohm volume control.
R_8 — 3300 ohms, ½ watt.
R_9, R_{12}, R_{19} — 0.47 megohm, ½ watt.
R_{11} — 0.15 megohm, ½ watt.
R_{14} — 10,000 ohms, 1 watt.
R_{15} — 2000-ohm wire-wound volume control.
R_{16} — 0.33 megohm, ½ watt.
R_{17} — 1500 ohms, ½ watt.
R_{18} — 82,000 ohms, ½ watt.
R_{20} — 150 ohms, 10 watts.
R_{21}, R_{22} — 39 ohms, 2 watts.
L_1, L_2, L_3 — 125 mh.
L_4 — 10 henrys, 60 ma.
L_5 — 10 henrys, 35 ma.
J_1 — 115-v. a.c. connector.
S_1, S_2 — S.p.s.t. toggle.
S_3 — D.p.d.t. toggle.
T_1 — Power transformer, 350 volts each side c.t., 70 ma.; 5 volts, 2 amp.; 6.3 volts, 3 amp. (Stancor P-4078).
T_2 — Output transformer, 5000 ohms (total primary) to line or voice coil.

tube from r.f. fields from the transmitter. The 6AL5 is between the 6SL7GT and the 6V6. The 6J5 is just to the rear of the 6V6, and the output transformer, T_2, is to its right. Along the front edge of the chassis are the microphone connector; gain control, R_7; clipper-filter switch, S_3; the "output" control, R_{15}; and — at the far right — the "B" voltage and a.c. toggle switches.

The low-pass filter is built as a unit on a 2 × 5-inch mounting board, as shown in Fig. 9-26. The coils are kept well separated and are mounted so that their axes are all at right angles. This prevents magnetic coupling between them, and is essential to good filter performance. In other respects the placement of parts in the filter is not critical. If the proper values of capacitance are not at hand, they can be made up by connecting smaller units in parallel. For example, a 0.01-μfd. paper and 0.005-μfd. mica can be paralleled to make 0.015 μfd. The filter unit occupies the upper right-hand corner in the bottom-view photograph, Fig. 9-27.

Particular care should be taken to reduce hum. The 6SJ7 grid lead must be shielded, and the heater wiring in the vicinity of the first two tubes should be kept in the corners of the chassis except where it is necessary to bring the ungrounded wire out to the socket terminal. It is worth while to try reversing the heater connections on the 6SJ7 to reduce hum. Reducing the gain at the lower frequencies also will reduce the hum in the output, and this may be done by decreasing the capacitance of C_5 and C_7 to 0.002 μfd. instead of the 0.01 μfd. specified.

The output transformer, T_2, in this unit is a low-impedance output type, with 500- and 200-ohm line taps as well as taps for a 'speaker voice coil. If the unit is close to the Class B modulator a Class B driver transformer can be substituted, if desired, or a 1-to-1 transformer can be used for grid-bias modulation.

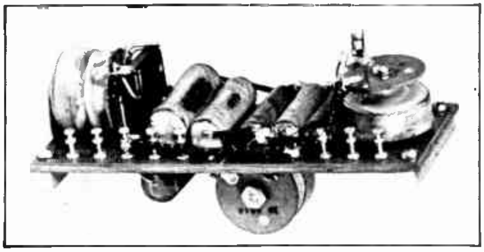


Fig. 9-26 — The low-pass filter is assembled as a unit on its own mounting board. Readily-available parts are used throughout.

Adjusting the Clipper-Filter Amplifier

The good effect of the low-pass filter in eliminating splatter can be entirely nullified if the amplifier stages following the filter can introduce appreciable distortion. That is a primary reason for the use of negative feedback in the output stage of the amplifier described. Amplifier stages following the unit must be operated well within their capabilities; in particular, the Class B output transformer (if a Class B modulator is to be driven) should be shunted by condensers to reduce the high-frequency response as described in the section on Class B modulators.

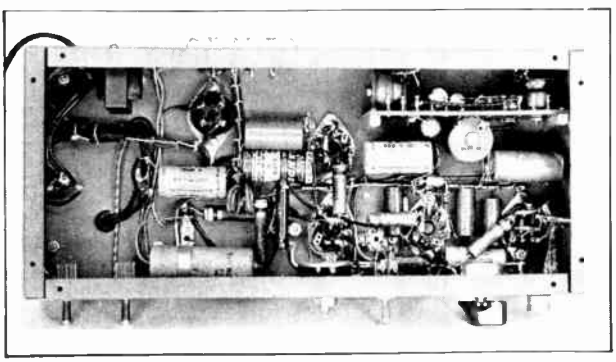
The setting of R_{15} is most important. It is most easily done with the aid of an oscilloscope (one having a linear sweep) and an audio oscillator, using the test set-up shown in the section on testing of speech equipment. Use a resistance load on the output transformer to reflect the proper load resistance (5000 ohms) at the plate of the 6V6. First set R_{15} at about ¼ the resistance from the ground end, switch in the clipper-filter, and apply a 500-cycle sine-wave signal to the microphone input. Increase the signal amplitude until clipping starts, as shown by flattening of both the negative and positive peaks of the wave. To check whether the clipping is taking place in the clipper or in the following amplifiers, throw S_3 to the "normal" or "out" position; the waveshape should return to normal. If it does not, return S_3 to the "in" position and reduce the setting of R_{15} until it does. Then reduce the amplifier gain by means of R_7 until the signal is just below the clipping level. At this point the signal should be a sine wave. Increase R_{15}, without touching R_7, until the wave starts to become distorted, and then back off R_{15} until distortion disappears.

Next, change the input-signal frequency to 2000 cycles, without changing the signal level. Slowly increase R_7 while observing the pattern. At this frequency it should be almost impossible to get anything except a sine wave through the filter, so if distortion appears it is the result of overloading in the amplifiers *following* the filter. Reduce the setting of R_{15} until the distortion disappears, even when R_7 is set at maximum and the maximum available signal from the audio oscillator is applied to the amplifier. The position of R_{15} should be marked at this point and the marked setting should never be exceeded.

To find the *operating* setting of R_{15}, leave the audio-oscillator signal amplitude at the value just under the clipping level and set up the complete transmitter for a modulation check, using the oscilloscope to give the trapezoidal pattern. With the Class C amplifier and modulator running, find the setting of R_{15} (keeping the audio signal just under the clipping level) that just gives 100-per-cent modulation. This setting should be below the maximum setting of R_{15} as previously determined; if it is not, the driver and modulator are not capable of modulating the transmitter 100 per cent and must be redesigned — or the Class C amplifier input must be lowered. Assuming a satisfactory setting is found, connect a microphone to the amplifier and set the amplifier gain control, R_7, so that the transmitter is modulated 100 per cent. Observe the pattern closely at different settings of R_7 to see if it is possible to overmodulate. If overmodulation does not occur at any setting of R_7, the transmitter is ready for operation and R_{15} may be locked in position; it need never be touched subsequently. If some overmodulation does occur, R_{15} should be backed off until it disappears and then locked.

In the absence of an oscilloscope the other methods of checking distortion described in the section on speech-amplifier testing may be used. The object is to prevent distortion in stages following the filter, so that when the clipping level is exceeded the following stages will be working within their capabilities.

Fig. 9-27 — Bottom view of the clipper-filter speech amplifier. Resistors and condensers are grouped around the sockets to which they connect.



6L6 Modulators for Low-Power Transmitters

Plate modulation for transmitters operating at final-stage plate power inputs up to 75 or 80 watts can be provided at relatively small cost by using Class AB 6L6s as modulators. The combined speech amplifier and modulator shown in Fig. 9-28 uses the 6L6s as Class AB_2 amplifiers and has an output (from the transformer secondary) of about 40 watts. The first stage is a 6SJ7 high-gain pentode amplifier,

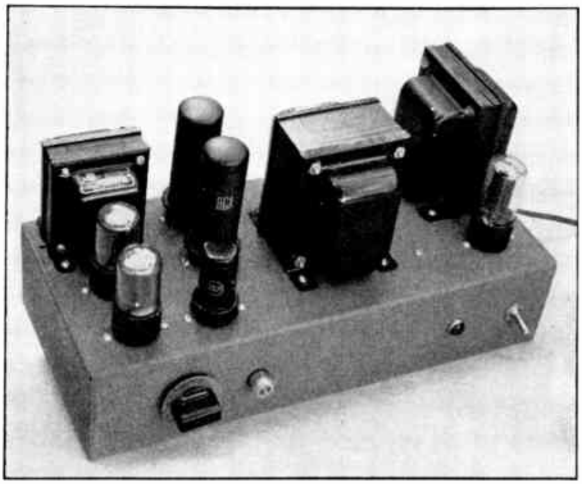


Fig. 9-28 — A 40-watt modulator of inexpensive construction. The second tube from the left, in the foreground, is the 6SJ7 first amplifier. The microphone connector is immediately below it on the chassis wall. Along the left edge, from the front, are the first and second 6SN7GTs and the driver transformer for the 6L6s. The output transformer is to the right of the 6L6s. The power transformer and rectifier are at the far right.

and is resistance coupled to one section of a 6SN7GT triode amplifier. The other section of the 6SN7GT is used as a single-tube phase inverter to obtain push-pull output. The grids of the push-pull 6L6s are driven by a 6SN7GT, with the two sections in push-pull, through transformer T_1. The gain control, R_6, is in the grid circuit of the first 6SN7GT section, and is shunted by condenser C_5 to reduce the high-frequency response. Condenser C_{11}, across the secondary of T_1, serves a similar purpose. The over-all circuit constants have been chosen so that the maximum response is in the most effective speech-frequency band. The response is down about 10 db. at 100 and 3000 cycles, as compared with the range 300–1500 cycles. The gain is more than sufficient for typical crystal microphones.

A power supply for the speech-amplifier stages and for the 6L6 heaters is included in the unit, but the power for the 6L6 plates and screens must be obtained from a separate supply. Fixed bias for the 6L6 grids is obtained from the built-in supply by taking the drop across R_{19}. This resistor, a potentiometer, should be adjusted so the voltage drop across it is 22.5 volts when the speech-amplifier stages are operating normally.

In building the amplifier, the usual precautions as to placement of components and wiring to avoid hum and feed-back should be observed. The microphone connector, J_1, should be located close to the 6SJ7 socket so the lead to the grid can be short. This lead also should be shielded.

The power supply for the 6L6s must have good voltage regulation, since the total current varies from approximately 95 ma. with no signal to 220 ma. at full output. A heavy-duty choke-input plate supply should be used; general design data will be found in Chapter Seven.

20-Watt Modulator

Fig. 9-31 is the circuit of a speech amplifier and modulator that has an output of approximately 20 watts. This circuit also uses 6L6s as output tubes, but the amplifier operates Class AB_1 and thus requires no driving power. Because of this, fewer voltage-amplifier stages are needed than in the case of the 40-watt amplifier. Push-pull input for the grids of the 6L6s is secured by using a single-plate-to-push-pull audio transformer between the 6J5 and the 6L6s. In this case it is economical to use a single power supply for the entire amplifier, so the low-voltage supply circuit shown in the 40-watt amplifier circuit may be omitted.

This amplifier can be used to plate-modulate

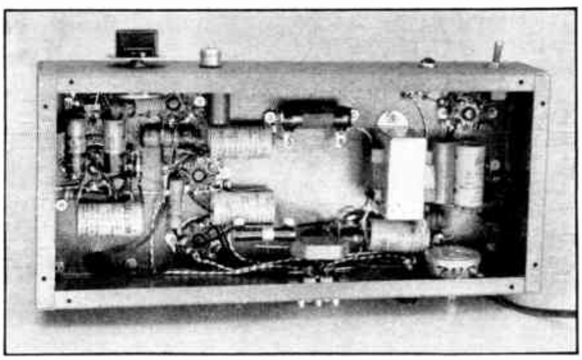


Fig. 9-29 — Underneath the chassis of the 40-watt modulator. The power-supply choke is mounted below chassis at the right. The bias-setting resistor, R_{19}, is on the rear chassis wall, at the lower right in this photograph. Other components are grouped near the tube socket with which they are associated.

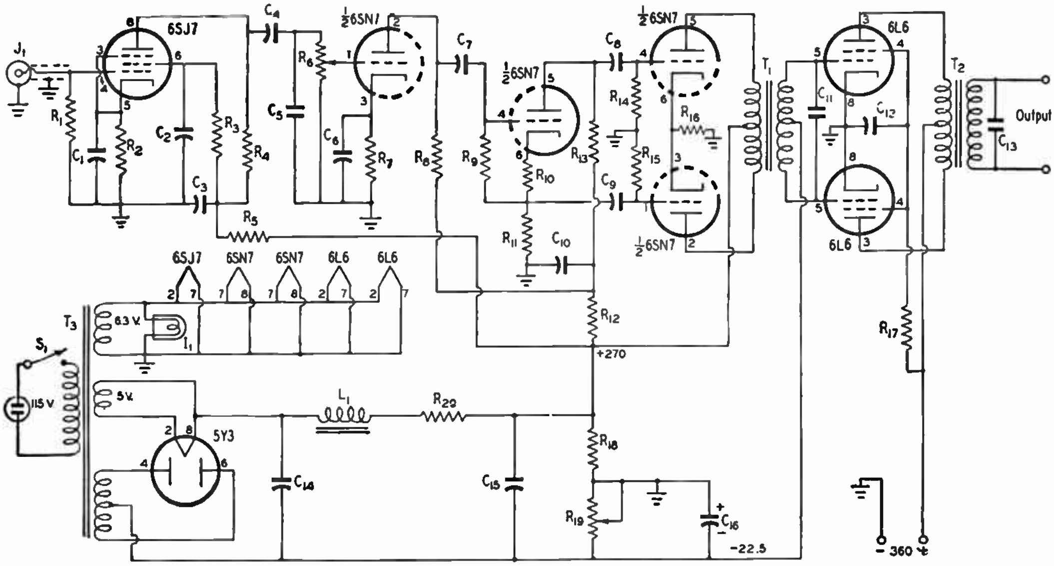


Fig. 9-30 — Circuit diagram of the 40-watt modulator.

C_1, C_6 — 25-μfd. 25-volt electrolytic.
C_2, C_4, C_7, C_8, C_9 — 0.1-μfd. 400-volt paper.
C_3, C_{10}, C_{12}, C_{14}, C_{15} — 8-μfd. 450-volt electrolytic.
C_5 — 470-μμfd. mica.
C_{11} — 0.1-μfd. 600-volt paper.
C_{13} — 0.01-μfd. 1200-volt mica.
C_{16} — 50-μfd. 50-volt electrolytic.
R_1 — 4.7 megohms, ½ watt.
R_2, R_7 — 1500 ohms, ½ watt.
R_3 — 1.5 megohms, ½ watt.
R_4 — 0.22 megohm, ½ watt.
R_5 — 47,000 ohms, ½ watt.
R_6 — 0.5-megohm potentiometer.
R_8, R_{13} — 56,000 ohms, ½ watt.
R_9, R_{14}, R_{15} — 0.47 megohm, ½ watt.
R_{10} — 18,000 ohms, ½ watt.
R_{11} — 39,000 ohms, ½ watt.
R_{12} — 10,000 ohms, 1 watt.
R_{16} — 470 ohms, 1 watt.
R_{17} — 7500 ohms, 10 watts.
R_{18} — 7000 ohms, 25 watts.
R_{19} — 1000-ohm wire-wound potentiometer, 4 watts.
R_{20} — 1200 ohms, 1 watt.
L_1 — Smoothing choke: 12 henrys, 80 ma. (Thordarson T20C53).
I_1 — 6.3-volt pilot lamp.
J_1 — Microphone-cable connector (Amphenol).
T_1 — Class AB2 driver transformer, p.p. plates to p.p. grids (Stancor A-4416).
T_2 — Modulation transformer, 3800 ohms to desired load (unit shown is Stancor A-3893).
T_3 — Power transformer: 350 volts each side center-tap, 70 ma.; 5 volts, 3 amp.; 6.3 volts, 3 amp. (Stancor P-4078).

an input of 40 watts to the r.f. amplifier. It is necessary, of course, to choose the proper output-transformer turns ratio to couple the modulator and modulated amplifier. The output stage is designed to work into a plate-to-plate load of 9000 ohms.

For the maximum power output of 20 watts, the plate supply for the amplifier must deliver 145 ma. at 360 volts. A condenser-input supply of ordinary design (Chapter Seven) may be used. The total plate current is approximately 120 ma. with no signal and 145 ma. at full output. If no more than 12 or 13 watts is needed, R_9 and R_{10} may be omitted and all tubes fed directly from a "B" supply giving approximately 175 ma. at 270 volts.

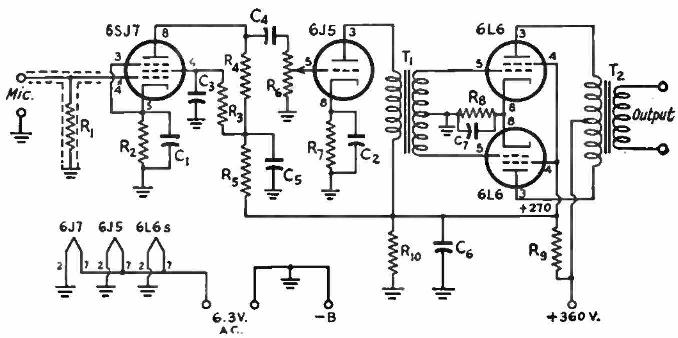


Fig. 9-31 — Circuit diagram of a low-cost modulator capable of power outputs up to 20 watts.

C_1, C_2 — 20-μfd. 50-volt electrolytic.
C_3 — 0.1-μfd. 200-volt paper.
C_4 — 0.01-μfd. 400-volt paper.
C_5, C_6 — 8-μfd. 450-volt electrolytic.
C_7 — 50-μfd. 50-volt electrolytic.
R_1 — 4.7 megohms, ½ watt.
R_2 — 1500 ohms, ½ watt.
R_3 — 1.5 megohms, ½ watt.
R_4 — 0.22 megohm, ½ watt.
R_5 — 47,000 ohms, ½ watt.
R_6 — 1-megohm volume control.
R_7 — 1500 ohms, 1 watt.
R_8 — 250 ohms, 10 watts.
R_9 — 2000 ohms, 10 watts.
R_{10} — 20,000 ohms, 25 watts.
T_1 — Interstage audio transformer, single plate to p.p. grids, ratio 3:1.
T_2 — Output transformer, type depending on requirements.

An 807 Modulator and Speech Amplifier

The combined speech amplifier and modulator unit shown in Fig. 9-32 is simple and inexpensive in design and, with the exception of the plate supply for the modulator tubes, is contained on a chassis measuring 3 × 8 × 17 inches. With a 750-volt plate supply, the push-pull Class AB_2 807s are capable of a tube output of 120 watts, or enough to plate-modulate a Class C stage with 200 watts input, allowing for moderate losses in the modulation transformer.

As shown in Fig. 9-33, the first tube in the speech amplifier is a 6J7 (a 6SJ7 may be substituted). A 6SN7GT is used in the second stage, one section serving as a voltage amplifier and the other as a phase inverter of the self-balancing type. The gain control for the amplifier is in the grid circuit of the first half of the tube. The third tube, also a 6SN7GT, is a push-pull amplifier, transformer-coupled to the grids of the 807s.

A power supply for the three tubes preceding the 807s is built on the same chassis. Voltage for the 807 screens is taken from this same supply. The negative return of the supply goes to the chassis through the adjustable arm of potentiometer R_{17}, which is connected in series with the bleeder resistor, R_{16}. The voltage developed in the section of R_{17} below the adjustable arm is negative with respect to chassis, and is used to provide fixed bias for the 807s. C_{11} is connected across this section of R_{17} to by-pass any a.f. current that might flow through the resistor. A separate filament transformer is provided for the 807 heaters, since the total heater power required by all the tubes in the amplifier is somewhat in excess of the rating of the 6.3-volt winding on the ordinary small power transformer.

Resistors R_{14} and R_{15} and condenser C_8 are placed in the 807 screen circuit to suppress the r.f. parasitic oscillations that sometimes occur with these tubes. Their use is principally a precautionary measure, and they may not be required in some installations.

The frequency response of this unit is maximum in the range from about 200 to 2500 cycles, for greatest voice effectiveness and minimum width of the r.f. channel. Frequencies above 2500 cycles are attenuated by condensers C_{12} and C_{13}, the former across the secondary of the driver transformer and the latter across the secondary of the output transformer. The capacitance values given are about optimum for the types of transformers specified and should be close to optimum for other transformers of similar ratings. The voltage rating of C_{13} should be at least equal to the d.c. voltage on the modulated r.f. amplifier.

The photographs show the general layout of components. The 6J7 and 6SN7GT phase inverter are in line at the left-hand front edge of the chassis. The 6SN7GT driver and 5Y3GT rectifier are to the rear of the phase inverter.

The bottom view shows the by-pass condensers and resistors grouped around the sockets to which they connect. The bias-control potentiometer, R_{17}, is mounted on the rear edge of the chassis. A jack shield (National JS-1) covers the microphone jack, and the first-stage grid resistor, R_1, is mounted inside this shield. The lead to the 6J7 grid cap must be shielded and the shield grounded.

The No. 1 terminals of the driver transformer specified should be connected to the grids of the 807s. If a different transformer is used, it should have a primary-to-secondary ratio (total) of about 1-to-1 to couple the

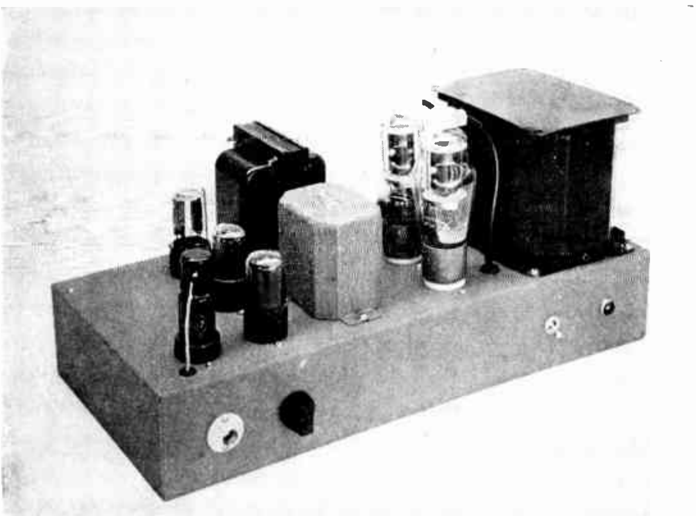


Fig. 9-32 — **A speech amplifier and 807 modulator for plate modulation of transmitters up to 200 watts input. The microphone jack and the gain control are at the left end of the chassis. The audio components and tubes occupy the front section, and the power supply for the driver tubes is laid out along the rear edge. The driver transformer is in the center foreground, with the power-supply transformer directly behind it. The large transformer at the right is the modulation transformer.**

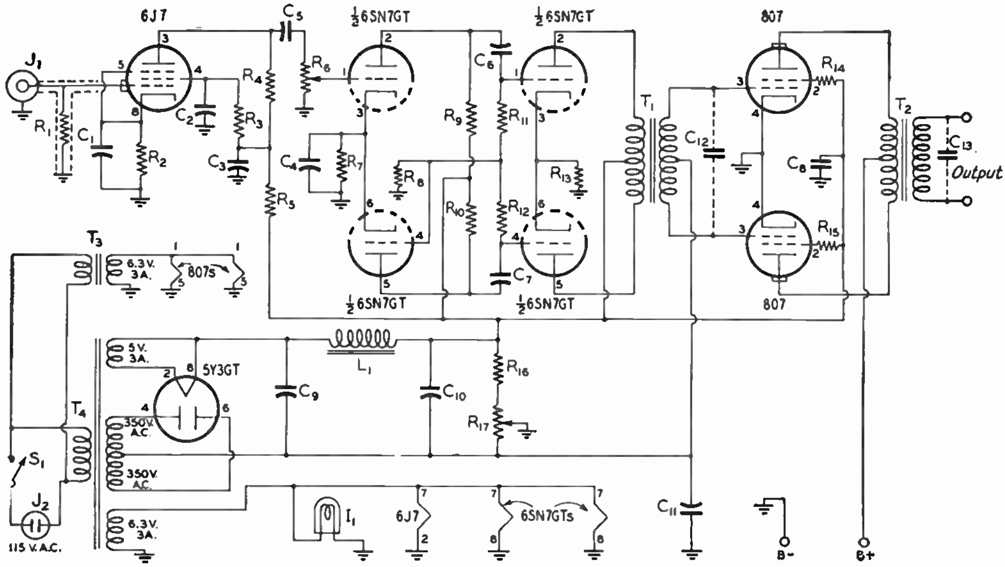


Fig. 9-33 — Circuit diagram of the push-pull 807 speech amplifier-modulator.

C_1 — 10-μfd. 50-volt electrolytic.
C_2 — 0.1-μfd. 400-volt paper.
C_3, C_9, C_{10} — 8-μfd. 450-volt electrolytic.
C_4, C_{11} — 50-μfd. 50-volt electrolytic.
C_5, C_6, C_7 — 0.01-μfd. 400-volt paper.
C_8 — 0.0068-μfd. mica.
C_{12} — 0.001-μfd. mica (see text).
C_{13} — 0.02-μfd. mica (see text).
R_1 — 1 megohm.
R_2, R_7 — 1500 ohms.
R_3 — 1.5 megohms.
R_4, R_8, R_{11}, R_{12} — 0.22 megohm.
R_5 — 47,000 ohms.
R_6 — 1-megohm volume control.
R_9, R_{10} — 0.1 megohm.
R_{13} — 470 ohms.
R_{14}, R_{15} — 100 ohms.
R_{16} — 15,000 ohms, 10 watts.
R_{17} — 1000-ohm wire-wound potentiometer.
(All resistors ½ watt unless otherwise noted.)
L_1 — Smoothing choke, 30 hy., 75 ma., 340-ohm d.c. resistance (Utah 4002).
I_1 — 6.3-volt a.c. pilot-lamp-and-socket assembly.
J_1 — Microphone-cable jack.
J_2 — Panel-mounting a.c. plug (Amphenol 61-M1).
S_1 — S.p.s.t. switch.
T_1 — Push-pull plates to push-pull grids (UTC S-9).
T_2 — Output transformer, type depending on requirements. A multitap transformer (UTC VM-3) is shown in photos.
T_3 — Filament transformer, 6.3 volts, 3 amp. (Thordarson T-21F10).
T_4 — Power transformer, 350 volts a.c. each side of center-tap, 70-ma. rating. Filament windings: 5 v., 3 amp.; 6.3 v., 3 amp. (Stancor P-4078).

6SN7GT and 807 grids properly. The output-transformer turns ratio will depend on the type of operation selected and the modulating impedance of the Class C amplifier. Operated at ICAS ratings, the 807s will deliver a tube output of 120 watts into a plate-to-plate load of 6950 ohms. This requires a plate supply capable of delivering 240 ma. at 750 volts. At CCS ratings the tubes will deliver 80 watts into a 6400-ohm load and require a 600-volt 200-ma. plate supply. The bias should be set, using R_{17}, to give −32 volts between the negative plate-supply terminal and chassis for ICAS operation, and to −30 volts for CCS operation.

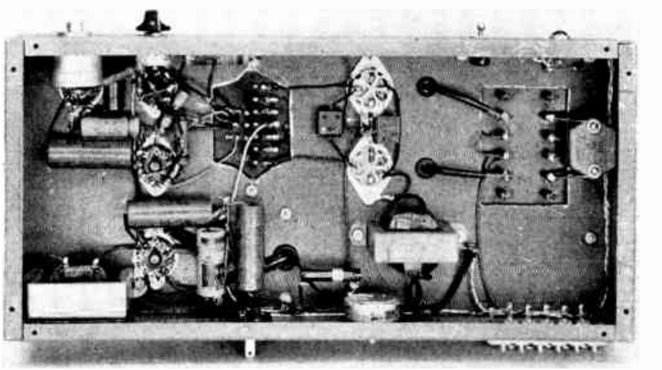


Fig. 9-34 — Below-chassis view of the 807 modulator. The shielded microphone jack is in the upper left-hand corner. The filter choke is mounted in the lower left-hand corner and the 807 filament transformer is to the rear and slightly to the right of the 807 tube sockets. The condenser for attenuating the high audio frequencies, shown at the right-hand end of the chassis, is supported by No. 12 wire leads which connect to the output terminals of the modulation transformer.

Class-B Modulators and Drivers

CLASS-B MODULATORS

Plate modulation of all but low-power transmitters requires so much audio power that the Class B amplifier is the only practical type to use. (Included in the Class B category are high-power modulators of the Class AB_2 type; whether the operation is in one class or the other is principally a matter of degree.)

Class B modulator circuits are practically identical no matter what the power output of the modulator. The diagrams of Fig. 9-35 therefore will serve for any modulator of this type that the amateur may elect to build. The triode circuit is given at A and the circuit for tetrodes at B. When small tubes with indirectly-heated cathodes are used, the cathodes should be connected to ground.

Modulator Tubes

Class B audio ratings of various types of transmitting tubes are given in the tube tables of Chapter Twenty-Five. Choose a pair of tubes that is capable of delivering sine-wave audio power equal to somewhat more than half the d.c. input to the modulated Class C amplifier. It is sometimes convenient to use tubes that will operate at the same plate voltage as that applied to the Class C stage, because one power supply of adequate current capacity may then suffice for both stages.

In estimating the output of the modulator, remember that the figures given in the tables are for the *tube* output only, and do not include output-transformer losses. To be adequate for modulating the transmitter, the modulator should have a theoretical power capability about 25 per cent greater than the actual power needed for modulation.

Matching to Load

In giving Class B ratings on power tubes, manufacturers specify the plate-to-plate load impedance into which the tubes must operate to deliver the rated audio power output. This load impedance seldom is the same as the modulating impedance of the Class C r.f. stage, so a match must be brought about by adjusting the turns ratio of the coupling transformer. The required turns ratio, primary to secondary, is

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

where N = Turns ratio, primary to secondary

Z_m = Modulating impedance of Class C r.f. amplifier

Z_p = Plate-to-plate load impedance for Class B tubes

Example: The modulated r.f. amplifier is to operate at 1250 volts and 250 ma. The power input is

$$P = EI = 1250 \times 0.25 = 312 \text{ watts}$$

so the modulating power required is 312/2 = 156 watts. Increasing this by 25% to allow for losses and a reasonable operating margin gives 156 × 1.25 = 195 watts. The modulating impedance of the Class C stage is

$$Z_m = \frac{E}{I} = \frac{1250}{0.25} = 5000 \text{ ohms.}$$

From the tube tables a pair of Class B tubes is selected that will give 200 watts output when working into a 6900-ohm load, plate-to-plate. The primary-to-secondary turns ratio of the modulation transformer therefore should be

$$N = \sqrt{\frac{Z_p}{Z_m}} = \sqrt{\frac{6900}{5000}} = \sqrt{1.38} = 1.175{:}1.$$

Commercial Class B output transformers usually are rated to work between specified primary and secondary impedances and frequently are designed for specific Class B tubes. In such a case, it will be unnecessary to calculate the turns ratio when the recommended tube combination is used. Many transformers are provided with primary and secondary taps, so that various turns ratios can be

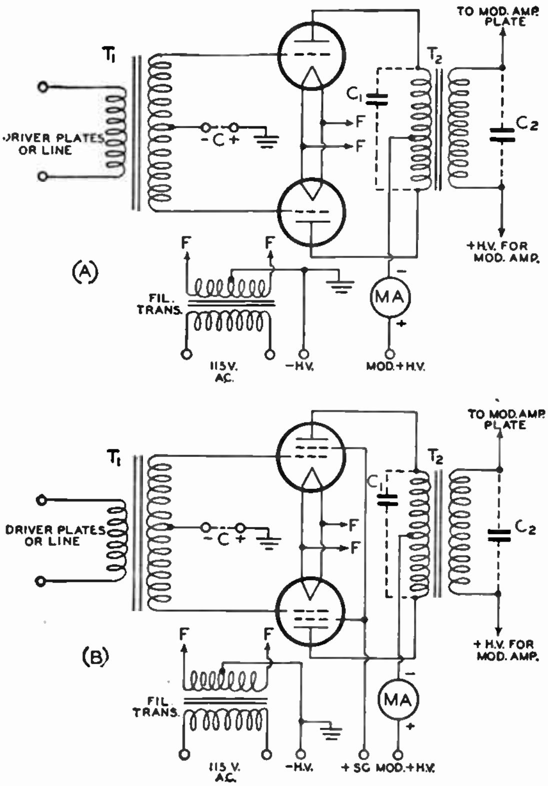


Fig. 9-35 — Class B modulator circuit diagrams. Tubes and circuit considerations are discussed in the text.

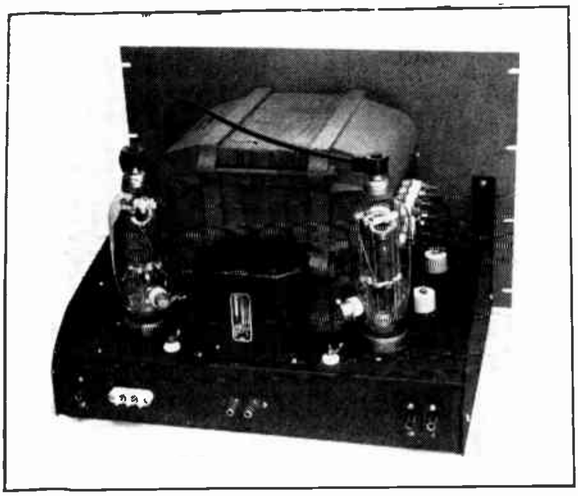


Fig. 9-36 — A typical chassis layout for a Class B modulator. Beyond adequate insulation for the voltages used, and sufficient ventilation for the modulator tubes, no particular constructional precautions are necessary. If the size of the components makes it necessary to use more than one chassis, the driver transformer may be included with the speech amplifier. In such case it is advisable to shield the "hot" audio leads to the modulator grids if they have to run any considerable distance.

obtained to meet the requirements of various tube combinations.

It may be that the exact turns ratio required by a particular tube combination cannot be secured, even with a tapped modulation transformer. *Small* departures from the proper turns ratio will have no serious effect if the modulator is operating well within its capabilities; if the actual turns ratio is within 10 per cent of the ideal value the system will operate satisfactorily. Where the discrepancy is larger, it is always possible to choose a new set of operating conditions for the Class C stage to give a modulating impedance that can be matched by the turns ratio of the available transformer. This may require operating the Class C amplifier at higher voltage and less plate current, if the modulating impedance must be increased, or at lower voltage and higher current if the modulating impedance must be decreased. However, this process cannot be carried too far without exceeding the ratings of the Class C tubes for either plate voltage or current, even though the power input is kept at the same figure. In such a case the only solution is to operate at reduced input and use less of the power available from the modulator.

Suppressing Audio Harmonics

Distortion in either the driver or Class B modulator itself will cause a.f. harmonics that may lie outside the frequency band needed for intelligible speech transmission. While it is almost impossible to avoid some distortion, it *is* possible to cut down the amplitude of the higher-frequency harmonics. The purpose of condensers C_1 and C_2 across the primary and secondary, respectively, of the Class B output transformer in Fig. 9-35 is to reduce the strength of harmonics and unnecessary high-frequency components existing in the modulation.

The condensers act with the leakage inductance of the transformer winding to form a rudimentary low-pass filter. The values of capacitance required will depend on the load resistance (modulating impedance of the Class C amplifier) and the leakage inductance of the particular transformer used. In general, capacitances between about 0.001 and 0.006 μfd. will be required; the larger values are necessary with the lower values of load resistance. A test set-up for measuring frequency response (described in a later section in this chapter) will quickly show the optimum values to use, if a small assortment of condensers is on hand for experimenting. The object is to find the combination of C_1 and C_2 that will give the most rapid reduction in response as the signal frequency is raised above about 2500 cycles.

The voltage rating of each condenser should at least be equal to the d.c. voltage at the transformer winding with which it is associated. In the case of C_2, part of the total capacitance required usually is supplied by the plate by-pass or blocking condenser of the modulated amplifier, so C_2 need only be large enough to make up the difference.

Grid Bias

Many modern transmitting tubes designed for Class B audio work can be operated without grid bias. Besides eliminating the need for a grid-bias supply, this reduces the variation in grid impedance over the audio-frequency cycle and thus gives the driver a more constant load into which to work. With these tubes, the grid return lead from the center-tap of the driver transformer secondary is simply connected to the filament center-tap or cathode.

When the tubes require bias, it should always be supplied from a *fixed* voltage source. Neither cathode bias nor grid-leak bias can be used with a Class B amplifier; with both types the bias changes with the amplitude of the signal voltage, whereas proper operation demands that the bias voltage be unvarying no matter what the strength of the signal. When only a small amount of bias is required

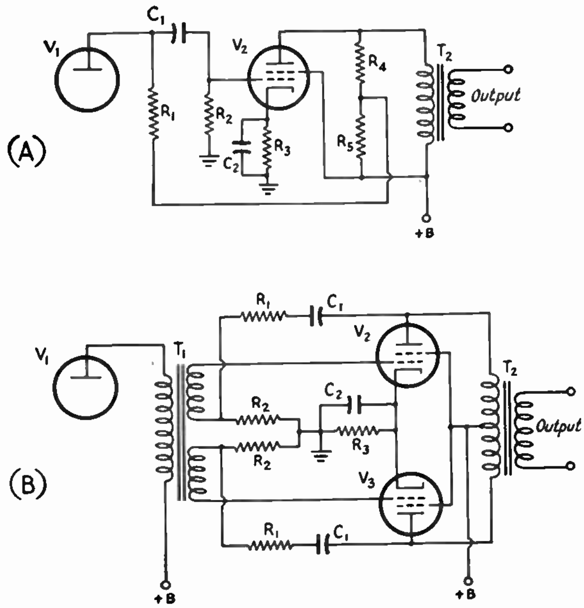


Fig. 9-38 — Negative feed-back circuits for drivers for Class B modulators. A — Single-ended beam-tetrode driver. If V_1 and V_2 are a 6J5 and 6V6, respectively, the following values are suggested: R_1, 47,000 ohms; R_2, 0.47 megohm; R_3, 250 ohms; R_4, R_5, 22,000 ohms; C_1, 0.01 μfd.; C_2, 50 μfd.

B — Push-pull beam-tetrode driver. If V_1 is a 6J5 and V_2 and V_3 6L6s, the following values are suggested: R_1, 0.1 megohm; R_2, 22,000 ohms; R_3, 250 ohms; C_1, 0.1 μfd.; C_2, 100 μfd.

Such high-frequency harmonies can be reduced by connecting condensers across both the primary and secondary of the output transformer as previously described.

Operation Without Load

Excitation should never be applied to a Class B modulator until after the Class C amplifier is turned on and is drawing the value of plate current required to present the rated load to the modulator. With no load to absorb the power, the primary impedance of the transformer rises to a high value and excessive audio voltages are developed across it — frequently high enough to break down the transformer insulation. If the modulator is to be tested separately from the transmitter, a resistance of the same value as the modulating impedance, and capable of dissipating the full power output of the modulator, should be connected across the transformer secondary.

● DRIVERS FOR CLASS-B MODULATORS

Class B amplifiers are driven into the grid-current region, so power is consumed in the grid circuit. The preceding stage or driver must be capable of supplying this power at the required peak audio-frequency grid-to-grid voltage. Both of these quantities are given in the manufacturer's tube ratings. The grids of the Class B tubes represent a variable load resistance over the audio-frequency cycle, because the grid current does not increase directly with the grid voltage. To prevent distortion, therefore, it is necessary to have a driving source that will maintain the waveform of the signal without distortion even though the load varies. That is, the driver stage must have good **regulation.** To this end, it should be capable of delivering somewhat more power than is consumed by the Class B grids, as previously described in the discussion on speech amplifiers. It is also desirable to use an input coupling transformer having a turns ratio giving the largest step-down in the voltage between the driver plate or plates and the Class B grids that will permit obtaining the specified grid-to-grid a.f. voltage.

The driver transformer, T or T_2 in Fig. 9-37, may couple directly between the driver tube and the modulator grids or may be designed to work into a low-impedance (200- or 500-ohm) line. In the latter case, a tube-to-line output transformer must be used at the output of the driver stage. This type of coupling is recommended only when the driver must be at a considerable distance from the modulator; the second transformer not only introduces additional losses but also impairs the voltage regulation of the driver stage.

Driver Tubes

The variation in grid resistance of a Class B amplifier over the audio-frequency cycle poses a special problem in the driver stage. To avoid distortion, the driver output *voltage* (not power) must stay constant (for a fixed signal voltage on its grid) regardless of the variations in load resistance.

The fundamental requirement for good voltage regulation in any electrical generator is that the internal resistance must be low. In a vacuum-tube amplifier, this means that the tubes must have a low value of plate resistance. The best tubes in this respect are low-μ triodes (the 6A3 is an example) and the worst are tetrodes and pentodes as represented by the 6V6 and 6L6. This does not mean that tetrodes (or pentodes) cannot be used, but it does mean that they should not be used without taking measures to reduce the effective plate resistance (see next section).

In selecting a driver stage always choose Class A or AB_1 operation in preference to Class AB_2. This not only simplifies the speech-amplifier design but also makes it easier to apply negative feed-back to tetrodes for reduction of plate resistance. It is possible to obtain a tube power output of approximately 25 watts (from 6L6s) without going beyond Class AB_1 operation; this is ample driving power for the popular Class B modulator tubes, even when a kilowatt transmitter is to be modulated.

The rated tube output (as shown by the

it can be obtained conveniently from a few dry cells. When greater values of bias are required, a heavy-duty "B" battery may be used if the grid current does not exceed 40 or 50 milliamperes on voice peaks. Even though the batteries are charged by the grid current rather than discharged, a battery will deteriorate with time and its internal resistance will increase. When the increase in internal resistance becomes appreciable, the battery tends to act like a grid-leak resistor and the bias varies with the applied signal. Batteries should be checked with a voltmeter occasionally while the amplifier is operating. If the bias varies more than 10 per cent or so with voice excitation the battery should be replaced.

As an alternative to batteries, a regulated bias supply may be used. This type of supply is described in Chapter Seven.

Plate Supply

The plate supply for a Class B modulator should be sufficiently well filtered to prevent hum modulation of the r.f. stage. An additional requirement is that the output condenser of the supply should have low reactance, at 100 cycles or less, compared to the load into which each tube is working. (This load is one-fourth of the plate-to-plate load resistance.) A 4-μfd output condenser with a 1000-volt supply, or a 2-μfd. condenser with a 2000-volt supply usually will be satisfactory. With other plate voltages, condenser values should be in inverse proportion to the plate voltage.

To keep distortion at a minimum, the voltage regulation of the plate supply should be as good as it can be made. If the d.c. output voltage of the supply varies with the amount of current taken, it should be kept in mind that the voltage at *maximum* current determines the amount of power that can be taken from the modulator without distortion. A supply whose voltage drops from 1500 at no load to 1250 at the full modulator plate current is a 1250-volt supply, so far as the modulator is concerned, and any estimate of the power output available should be based on the lower figure.

It is particularly important, in the case of a tetrode Class B stage, that the screen-voltage power-supply source have excellent regulation, to prevent distortion. The screen voltage should be set as exactly as possible to the recommended value for the tube.

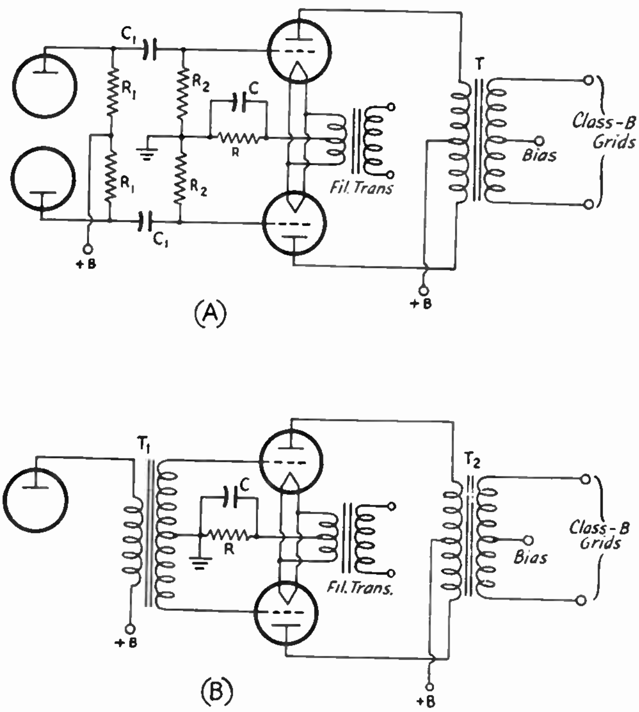


Fig. 9-37 — Triode driver circuits for Class B modulators. A, resistance coupling to grids; B, transformer coupling. R_1 in A is the plate resistor for the preceding stage, value determined by the type of tube and operating conditions as given in Table 9-I. C_1 and R_2 are the coupling condenser and grid resistor, respectively: values also may be taken from Table 9-I.
In both circuits the output transformer, T, T_2, should have the proper turns ratio to couple between the driver tubes and the Class B grids. T_1 in B is usually a 2:1 transformer, secondary to primary. R, the cathode resistor, should be calculated for the particular tubes used. The value of C, the cathode by-pass, is determined as described in the text.

Overexcitation

When a Class B amplifier is overdriven in an attempt to secure more than the rated power, distortion increases rapidly. The high-frequency harmonics which result from the distortion modulate the transmitter, producing spurious sidebands which can cause serious interference over a band of frequencies several times the channel width required for speech. This may happen even though the transmitter is not being overmodulated. It *will* happen if the modulator is incapable of delivering the power required to modulate the transmitter fully, or if the Class C amplifier is not adjusted to give the proper modulating impedance.

As previously stated, the tubes used in the Class B modulator should be capable of somewhat more than the power output nominally required. In addition, the Class C amplifier should be adjusted to give the proper modulating impedance and the correct output transformer turns ratio should be used. Even though means may be incorporated in the speech amplifier to attenuate frequencies above those necessary for intelligible speech, it is still possible for high-frequency sidebands to be radiated if distortion occurs in the modulator, or if the transmitter is overmodulated.

tube tables) should be reduced by about 20 per cent to allow for losses in the Class B input transformer. If two transformers are used, tube-to-line and line-to-grids, allow about 35 per cent for transformer losses. Another 25 per cent should be allowed, if possible, as a safety factor and to improve the voltage regulation.

Fig. 9-37 shows representative circuits for a push-pull triode driver using cathode bias. If the amplifier operates Class A, the cathode resistor need not be by-passed, because the a.f. currents from each tube flowing in the cathode resistor are out of phase and cancel each other. However, in Class AB operation this is not true; considerable distortion will be generated at high signal levels if the cathode resistor is not by-passed. The by-pass capacitance required can be calculated by a simple rule: the cathode resistance in ohms multiplied by the by-pass capacitance in microfarads should equal at least 25,000. The voltage rating of the condenser should be equal to the maximum bias voltage. This can be found from the maximum-signal plate current and the cathode resistance.

Example: A pair of 6A3s is to be used in Class AB_1, self-biased. From the tube tables, the cathode resistance should be 780 ohms and the maximum-signal plate current 120 ma. From Ohm's Law,

$$E = RI = 780 \times 0.12 = 93.6 \text{ volts}$$

From the rule mentioned previously, the by-pass capacitance required is

$$C = 25{,}000/R = 25{,}000/780 = 32 \mu\text{fd.}$$

A 40- or 50-μfd. 100-volt electrolytic condenser would be satisfactory.

Negative Feed-Back

Whenever tetrodes or pentodes are used as drivers for Class B modulators, negative feed-back should be used in the driver stage. This will reduce the distortion caused by the variable load resistance represented by the Class B grids. It also reduces the distortion inherent in the driver stage itself, when properly applied. The effect of feed-back is to reduce the apparent plate resistance of the driver, and this in turn helps to maintain the a.f. output voltage at a more constant level (for a constant signal on the grid) when the load resistance varies. It is readily possible to reduce the plate resistance to a value comparable to or lower than that of low-μ triodes such as the 6A3.

Suitable circuits for single-ended and push-pull tetrodes are shown in Fig. 9-38. Fig. 9-38A shows resistance coupling between the preceding stage and a single tetrode, such as the 6V6, that operates at the same plate voltage as the preceding stage. Part of the a.f. voltage across the primary of the output transformer is fed back to the grid of the tetrode, V_2, through the plate resistor of the preceding tube, V_1. The amount of voltage so fed back is determined by the voltage divider, R_4R_5. The total resistance of R_4 and R_5 in series should be large compared to the rated load resistance of V_2. Instead of the voltage divider, a tap on the transformer primary can be used to supply the feed-back voltage, if such a tap is available.

The amount of feed-back voltage that appears *at the grid* of tube V_2 is determined by R_1, R_2 and the plate resistance of V_1, as well as by the relationship between R_4 and R_5. Calculation of the feed-back voltage, although not mathematically difficult, is not ordinarily practicable because the plate resistance of V_1 is seldom known at the particular operating conditions used. Circuit values for a typical tube combination are given in detail in Fig. 9-38.

The push-pull circuit in Fig. 9-38B requires an audio transformer with a split secondary. The feed-back voltage is obtained from the plate of each output tube by means of the voltage divider, R_1R_2. The blocking condenser, C_1, prevents the d.c. plate voltage from being applied to R_1R_2; the reactance of this condenser should be low, compared with the sum of R_1 and R_2, at the lowest audio frequency to be amplified. Also, the sum of R_1 and R_2 should be high compared with the rated load resistance for V_2 and V_3.

In this circuit the feed-back voltage that is developed across R_2 also appears at the grid of V_2 (or V_3) because there is no appreciable current flow (in the usual audio range) through the transformer secondary and grid-cathode circuit of the tube, provided the tubes are not driven to grid current. If the grid-cathode impedance of the tubes is relatively low, as it is when grid current flows, the feed-back voltage decreases because of the voltage drop through the transformer secondary. The circuit should not be used with tubes that are operated Class

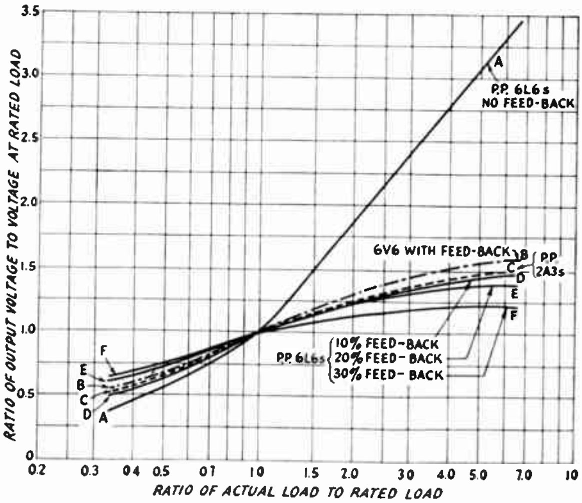


Fig. 9-39 — Output voltage regulation of two types of beam-tetrode drivers with negative feed-back. For comparison, the regulation with a pair of 2A3s (no feed-back) also is shown.

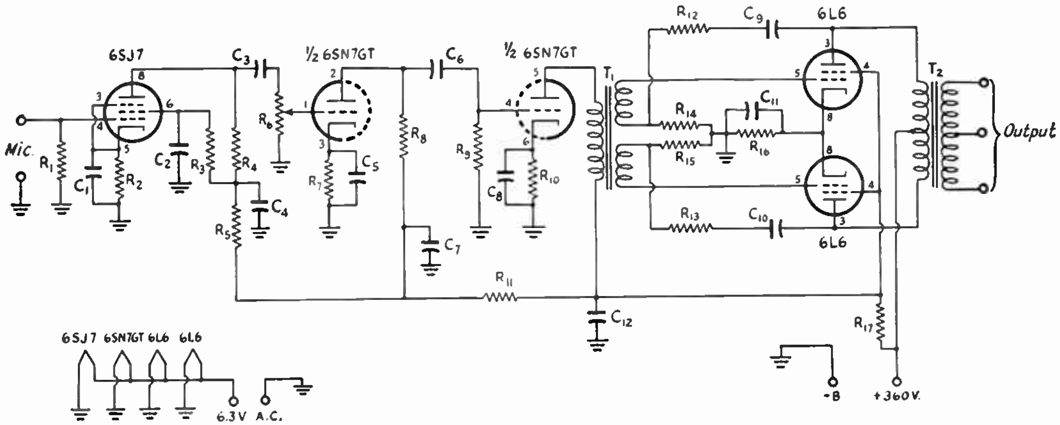


Fig. 9-40 — Circuit diagram of speech amplifier using 6L6s with negative feed-back, suitable for driving Class B modulators up to 500 watts output.

C_1, C_5, C_8 — 20-μfd. 25-volt electrolytic.
C_2, C_9, C_{10} — 0.1-μfd. 400-volt paper.
C_3, C_6 — 0.01-μfd. 600-volt paper.
C_4, C_7, C_{12} — 10-μfd. 450-volt electrolytic.
C_{11} — 100-μfd. 50-volt electrolytic.
R_1 — 2.2 megohms, ½ watt.
R_2, R_7 — 1500 ohms, ½ watt.
R_3 — 1.5 megohms, ½ watt.
R_4 — 0.22 megohm, ½ watt.
R_5, R_8 — 47,000 ohms, ½ watt.
R_6 — 1-megohm volume control.
R_9 — 0.47 megohm, ½ watt.
R_{10} — 1500 ohms, 1 watt.
R_{11} — 10,000 ohms, ½ watt.
R_{12}, R_{13} — 0.1 megohm, 1 watt.
R_{14}, R_{15} — 22,000 ohms, ½ watt.
R_{16} — 250 ohms, 10 watts.
R_{17} — 2000 ohms, 10 watts.
T_1 — Interstage audio, 2:1 secondary (total) to primary, with split secondary winding.
T_2 — Class B input transformer to suit modulator tubes.

AB_2. The per cent feed-back is

$$n = \frac{R_2}{R_1 + R_2} \times 100$$

where n is the feed-back percentage, and R_1 and R_2 are connected as shown in the diagram. The higher the feed-back percentage, the lower the effective plate resistance. However, if the percentage is made too high the preceding tube, V_1, may not be able to develop enough voltage, through T_1, to drive the push-pull stage to maximum output without itself generating harmonic distortion. Distortion in V_1 is not compensated for by the feed-back circuit. If V_2 and V_3 are 6L6s operated self-biased in Class AB_1 with a load resistance of 9000 ohms, V_1 is a 6J5, and T_1 has a turns ratio of 2-to-1, total secondary to primary, it is possible to use over 30-per-cent feed-back without going beyond the output-voltage capabilities of the 6J5. Actually, it is unnecessary to use more than about 20-per-cent feed-back. This value reduces the effective plate resistance to the point where the output voltage regulation is better than that of 6A3s or 2A3s without feed-back.

Instead of the voltage-divider arrangement shown in Fig. 9-38B for obtaining feed-back voltage, a separate winding on the output transformer can be used, provided it has the proper number of turns to give the desired feed-back percentage. Special transformers are available for this purpose.

The improvement in constancy of output voltage resulting from the use of negative feed-back is shown graphically in Fig. 9-39. In order to compare the various types of tubes, the variation in output voltage is shown as a percentage of the output voltage when the tubes are working into the rated load. The load resistance also is expressed as a percentage of the rated load resistance for the particular tube, or pair of tubes, used.

SPEECH-AMPLIFIER CIRCUIT WITH NEGATIVE FEED-BACK

A circuit for a speech amplifier suitable for driving a Class B modulator is given in Fig. 9-40. In this amplifier the 6L6s are operated Class AB_1 and will deliver up to 20 watts to the grids of the Class B amplifier. The feed-back circuit requires no adjustment, but does require an interstage transformer with two separate secondary windings (split secondary).

This amplifier may be constructed along the same lines as in Fig. 9-28, observing the same precautions with respect to shielding the 6SJ7 grid circuit. Although the power output is the same as from the amplifier of Fig. 9-31, an additional voltage-amplifier stage is incorporated in the circuit. This is necessary because the voltage fed back from the plates to the grids of the 6L6s opposes the voltage from the preceding stage, so the latter must be increased in order to maintain the same power output from the 6L6s. In turn, this necessitates more over-all voltage gain than is required to drive Class AB_1 p.p. 6L6s without feed-back.

The output transformer, T_2, should be selected to work between a 9000-ohm plate-to-plate load and the grids of whatever Class B tubes will be used. The power-supply requirements for this amplifier are essentially the same as for the amplifier of Fig. 9-31.

Checking 'Phone-Transmitter Operation

SPEECH EQUIPMENT

Every 'phone transmitter requires checking before it is initially put on the air. An adequate job can be done with equipment that is neither elaborate nor expensive. A simple set-up is shown in Fig. 9-41. The only equipment that is not likely to be already at hand is the audio oscillator (the construction of a very simple one is described in Chapter Sixteen). The voltmeter — one that operates at audio frequencies is necessary — can be any multirange volt-ohm-milliammeter that has a rectifier-type a.c. range. The headset is included for aural checking of the amplifier performance.

The audio oscillator usually will have an output control, but if the maximum output voltage is in excess of a volt or so the output setting may be rather critical when a high-gain speech amplifier is being tested. In such cases an attenuator such as is shown in Fig. 9-41 is a convenience. Each of the two voltage dividers reduces the voltage by a factor of roughly 10 to 1, so that the over-all attenuation is about 100 to 1. The relatively low value of resistance, R_4, across the input terminals of the amplifier also will minimize stray hum pick-up on the connecting leads.

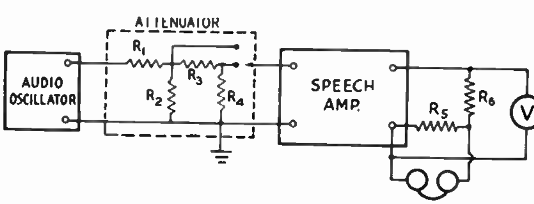


Fig. 9-41 — Simple test set-up for checking a speech amplifier. The audio-oscillator frequency range should be from about 100 to 5000 or more cycles. It is not necessary that it be continuously variable; a number of "spot" frequencies will be satisfactory. Suitable resistor values are: R_1 and R_3, 10,000 ohms; R_2 and R_4, 1000 ohms; R_6, rated load resistance for amplifier output stage; R_5, determine by trial for comfortable headphone level (25 to 100 ohms, ordinarily). V is a high-resistance a.c. voltmeter, multirange rectifier type.

As a preliminary check, cover the microphone input terminals with a metal shield (with the audio oscillator and attenuator disconnected) and, while listening in the headset, note the hum level with the amplifier gain control in the off position. The hum should be very low under these conditions. Then increase the gain-control setting to maximum and observe the hum; it will no doubt increase. Then connect the audio oscillator and attenuator and, starting from minimum signal, increase the audio input voltage until the voltmeter indicates full power output. (The voltage should equal $\sqrt{PR}$, where P is the expected power output in watts and R is the load resistance — R_6 in the diagram.) While increasing the input, listen carefully to the tone to see if there is any change in its character. When it begins to sound like a musical octave instead of a single tone, distortion is beginning. Assuming that the output is substantially without audible distortion at full output, substitute the microphone for the audio oscillator and speak into it in a normal tone while watching the voltmeter. Reduce the gain-control setting until the meter "kicks" nearly up to the full-power reading on voice peaks. Note the hum level, as read on the voltmeter, at this point; the hum level should not exceed one or two per cent of the voltage at full output.

If the hum level is too high, the amplifier stage that is causing the trouble can be located by temporarily short-circuiting the grid of each tube, in turn, to ground. When shorting a particular grid makes a marked decrease in hum, the hum presumably is coming from a *preceding* stage, although it is possible that it is getting its start in that particular grid circuit. If shorting a grid does *not* decrease the hum, the hum is originating either in the plate circuit of that tube or the grid circuit of the next. Aside from wiring errors, a defective tube, or inadequate plate-supply filtering, objectionable hum usually originates in the first stage of the amplifier.

If distortion occurs below the point at which the expected power output is secured, the stage in which it is occurring can be located by working from the last stage toward the front end of the amplifier, applying a signal to each grid in turn from the audio oscillator and adjusting the signal voltage for maximum output. In the case of push-pull stages, the signal may be applied to the primary of the interstage transformer — *after* disconnecting it from the plate-voltage source. Assuming that normal design principles have been followed and that all stages are theoretically working within their capabilities, the probable causes of distortion are wiring errors (such as accidental short-circuit of a cathode resistor), defective components, or use of wrong values of resistance in cathode and plate circuits.

Using the Oscilloscope

Speech-amplifier checking is facilitated considerably if an oscilloscope of the type having amplifiers and a linear sweep circuit is available. A typical set-up for using the oscilloscope is shown in Fig. 9-42. With the connections shown, the sweep circuit is not required but horizontal and vertical amplifiers are necessary. Audio voltage from the oscillator is fed directly to one oscilloscope amplifier (horizontal in this case) and the output of the speech amplifier is connected to the other. The 'scope amplifier gains should be adjusted so that each signal gives the same line length with the other signal shut off.

Under these conditions, when the input and output signals are applied simultaneously they are compared directly. If the speech amplifier is distortion-free and introduces no phase shift, the resulting pattern is simply a straight line, as shown at the upper left in Fig. 9-43, making an angle of about 45 degrees with the horizontal and vertical axes. If there is no distortion but there is some phase shift, the pattern will be a smooth ellipse, as shown at the upper right. The greater the phase shift the greater the tendency of the ellipse to grow into a circle. When there is even-harmonic distortion in the amplifier one end of the line or ellipse becomes curved, as shown in the second row in Fig. 9-43. With odd-harmonic distortion such as is characteristic of overdriven push-pull stages, the line or ellipse is curved at both ends.

Patterns such as these will be obtained when the input signal is a fairly good sine wave. They will tend to become complicated if the input waveform is complex and the speech amplifier introduces improper phase shifts. Most amplifiers will be quite satisfactory in this respect in the medium audio-frequency range, so it is advisable to check for distortion with a frequency in the vicinity of 500 cycles.

Generally speaking, it is easier to detect small amounts of distortion with the type of pattern shown in Fig. 9-43 than it is with the waveform pattern obtained by feeding the output signal to the vertical plates and making use of the linear sweep in the 'scope. This is because it is quite easy to determine whether or not a line is straight, but not so easy to decide whether or not a pattern displayed by the sweep circuits meets given specifications. The waveform pattern can be used satisfactorily, however, if the signal from the audio oscillator is a reasonably good sine wave. One simple method is to examine the output of the oscillator alone and trace the pattern on a sheet of transparent paper. The pattern given by the output of the amplifier can then be compared with the "standard" pattern by adjusting the oscilloscope gain to make the two patterns coincide as closely as possible. The pattern discrepancies are a measure of the distortion.

In using the oscilloscope care must be used to avoid introducing hum voltages that will upset the measurements. Hum pick-up on the 'scope leads or other exposed parts such as the amplifier load resistor or the voltmeter can be detected by shutting off the audio oscillator and speech amplifier and connecting first one and then the other to the vertical plates of the 'scope, setting the internal horizontal sweep to an appropriate width. The trace should be a straight horizontal line when the vertical gain control is set at the position used in the actual measurements. Waviness in the line indicates hum. If the hum is not in the 'scope itself (check by disconnecting the leads at the instrument) make sure that there is a good ground connection on all the equipment and, if necessary, shield the hot leads.

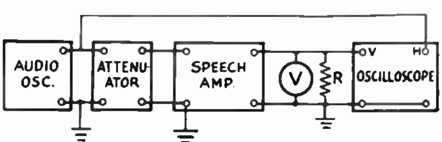


Fig. 9-12 — **Test set-up using the oscilloscope to check for distortion. These connections will result in the type of pattern shown in Fig. 9-13, the horizontal sweep being provided by the audio input signal. For waveform patterns, omit the connection between the audio oscillator and the horizontal amplifier in the 'scope, and use the horizontal linear sweep.**

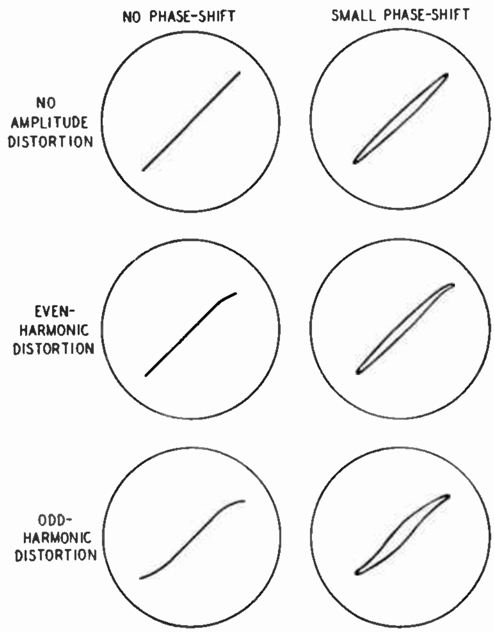


Fig. 9-13 — **Typical patterns obtained with the connections shown in Fig. 9-12. Depending on the number of stages in the amplifier, the pattern may slope upward to the right, as shown, or upward to the left. Also, depending on where the distortion originates, the curvature in the second row may appear either at the top or bottom of the line or ellipse.**

The oscilloscope can be used to good advantage in stage-by-stage testing to check waveforms at the grid and plate of each stage and thus to determine rapidly where a source of trouble may be located. When the 'scope is connected to circuits that are not at ground potential for d.c., a condenser of about 0.1 ufd. should be connected in series with the hot oscilloscope lead. The probe lead should be shielded so that it will not pick up hum.

CLASS-B MODULATORS

Once the speech amplifier is in satisfactory working condition, the Class B modulator can be checked by similar means. A simple circuit is shown in Fig. 9-44. The resistance of R_1 should be equal to the modulating impedance of the Class C amplifier to be modulated, and the resistor should have a power rating equal to the rated power output of the modulator. Calculate the voltage to be expected across R_1 at full output; if it exceeds the range of the meter the meter may be connected across say half or one-fourth of R_1 and the readings multiplied by 2 or 4, respectively. Only a few ohms will be needed at R_2, in the average case, to give a good signal in the headphones. As a safety precaution, ground the output terminal to which the headphones are connected and use a resistor at R_2 that has ample current-carrying capacity.

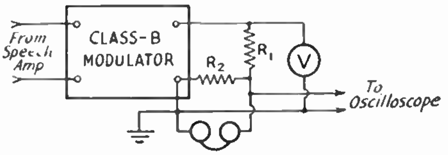


Fig. 9-44 — Set-up for checking a Class B modulator.

Hum will seldom be a problem in the modulator. Distortion may be checked as described previously; the oscilloscope is excellent for this purpose. If a variable-frequency audio oscillator is used, a check on the frequency response of the over-all system can be obtained by varying the oscillator frequency (check its output voltage at each frequency change) and observing the variation in the modulator output voltage. The high-frequency response of the system can be attenuated by trying condensers of various values across the primary and secondary of the output transformer, as pointed out in the discussion on Class B modulators. The object is to reduce the response above 3000 cycles to a low value as compared with the response in the 200- to 2500-cycle region, so that the channel occupied by the transmitter will not be excessive. A simple method of adjustment is to apply an audio tone of about 1500 cycles and increase its amplitude until distortion becomes noticeable; when this occurs the tone no longer sounds pure but sounds like a musical octave. The condenser values should then be adjusted until the test tone sounds pure again at the same signal amplitude.

THE MODULATED AMPLIFIER

Proper adjustment of a 'phone transmitter is aided immeasurably by the oscilloscope; it will give more information, more accurately, than almost any collection of other instruments that might be named. Furthermore, an oscilloscope that is entirely satisfactory for the purpose is not necessarily an expensive instrument; the cathode-ray tube and its power supply are about all that are needed. Amplifiers and linear sweep circuits are by no means necessary.

When using the tube without a sweep circuit, radio-frequency voltage from the modulated amplifier is applied directly to the vertical deflection plates of the tube, and audio-frequency voltage from the modulator is applied to the horizontal deflection plates. As the amplitude of the horizontal signal varies, the r.f. output of the transmitter also varies, and this produces a wedge-shaped pattern or **trapezoid** on the screen. If the oscilloscope has a horizontal sweep, the r.f. voltage is applied to the vertical plates as before (never through an amplifier) and the sweep produces a pattern that follows the modulation envelope of the transmitter output, provided the sweep frequency is lower than the modulation frequency. This produces a **wave-envelope** modulation pattern.

Oscilloscope connections for both types of patterns are shown in Fig. 9-45. The connections for the wave-envelope pattern are somewhat simpler than those for the trapezoidal figure. The vertical deflection plates are coupled to the amplifier tank coil (or an antenna coil) through a twisted-pair line and pick-up coil. As shown in the alternative drawing, a resonant circuit tuned to the operating frequency may be connected to the vertical plates, using link coupling between it and the transmitter. This will eliminate r.f. harmonics,

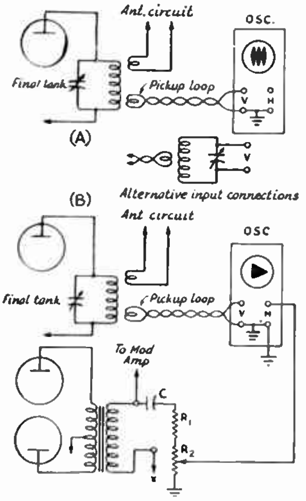


Fig. 9-45 — Methods of connecting an oscilloscope to the modulated r.f. amplifier for checking modulation. See text for discussion.

and the tuning control provides a means for adjustment of the pattern height.

To get a wave-envelope pattern the position of the pick-up coil should be varied until a carrier pattern, Fig. 9-46B, of suitable height is obtained. The horizontal sweep voltage should be adjusted to make the width of the pattern somewhat more than half the diameter of the screen. When voice modulation is applied, a rapidly-changing pattern of varying height will be obtained. When the maximum height of this pattern is just twice that of the carrier alone, the wave is being modulated

100 per cent. This is illustrated by Fig. 9-46D, where the point X represents the sweep line (reference line) alone, YZ is the carrier height, and PQ is the maximum height of the modulated wave. If the height is greater than the distance PQ, as illustrated in E, the wave is overmodulated in the upward direction. Overmodulation in the downward direction is indicated by a gap in the pattern at the reference axis, where a single bright line appears on the screen. Overmodulation in either direction may take place even when the modulation in the other direction is less than 100 per cent.

Connections for the trapezoidal pattern are shown in Fig. 9-45B. The vertical plates are coupled to the transmitter tank circuit

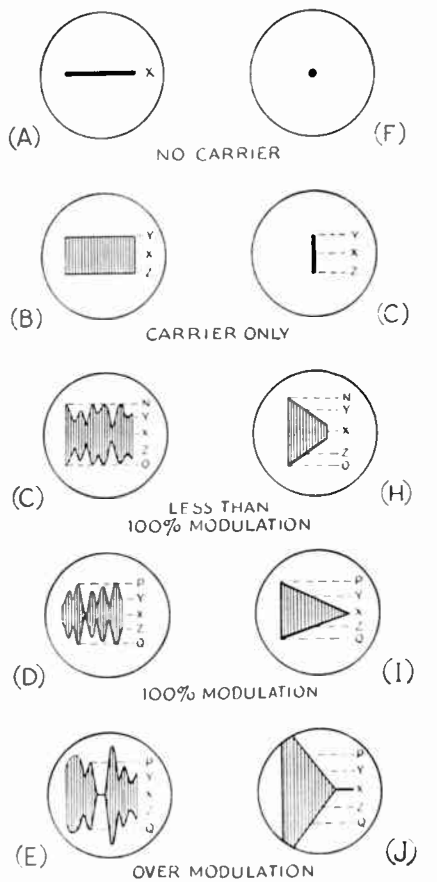


Fig. 9-46 — Wave-envelope and trapezoidal patterns representing different conditions of modulation.

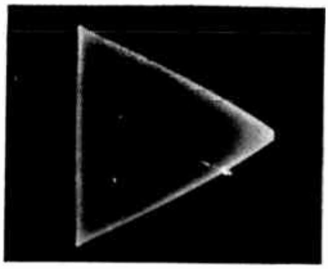
through a pick-up loop; alternatively, the tuned input circuit to the oscilloscope may be used. The horizontal plates are coupled to the output of the modulator through a voltage divider, R_1R_2. R_2 should be a potentiometer so the audio voltage can be adjusted to give a satisfactory horizontal sweep on the screen. R_2 may be a 0.25-megohm volume control. The value of R_1 will depend upon the audio output voltage of the modulator. This voltage is equal to $\sqrt{PR}$, where P is the audio power output of the modulator and R is the modulating impedance of the modulated r.f. amplifier. In the case of grid-bias modulation with a 1:1 output transformer, it will be satisfactory to assume that the a.c. output voltage of the modulator is equal to $0.7E$ for a single tube, or to $1.4E$ for a push-pull stage, where E is the d.c. plate voltage on the modulator. If the transformer ratio is other than 1:1, the voltage so calculated should be multiplied by the actual secondary-to-primary turns ratio.

The total resistance of R_1 and R_2 in series should be 0.25 megohm for every 150 volts of modulator output; for example, if the modulator output voltage is 600, the total resistance should be four (600/150) times 0.25 megohm, or 1 megohm. Then, with 0.25 megohm at R_2, R_1 should be 0.75 megohm. For good low-frequency coupling the capacitance, in microfarads, of the blocking condenser, C, should at least equal $0.004/R$, where R is the total resistance $(R_1 + R_2)$ in megohms. Thus in the example above, where R is 1 megohm, the capacitance required is 0.004 μfd. The voltage rating of the condenser should be at least twice the d.c. voltage applied to the modulated amplifier — that is, the same as the rating of the plate by-pass condenser in the final stage. The capacitance can be made up of two or more similar units in series, so long as the total capacitance is equal to that required, in case units of sufficient voltage rating are not available; or of two or more units in parallel if condensers having adquate voltage rating but insufficient capacitance are available.

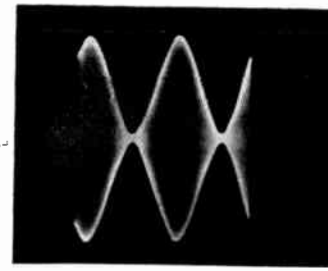
Trapezoidal patterns for various conditions of modulation are shown in Fig. 9-46 at F to J, each alongside the corresponding wave-envelope pattern. With no signal, only the cathode-ray spot appears on the screen. When the unmodulated carrier is applied, a vertical line appears; the length of the line should be adjusted, by means of the pick-up coil coupling, to a convenient value. When the carrier is modulated, the wedge-shaped pattern appears; the higher the modulation percentage, the wider and more pointed the wedge becomes. At 100-per-cent modulation it just makes a point on the axis, X, at one end, and the height, PQ, at the other end is equal to twice the carrier height, YZ. Overmodulation in the upward direction is indicated by increased height over PQ, and in the downward direction by an extension along the axis X at the pointed end

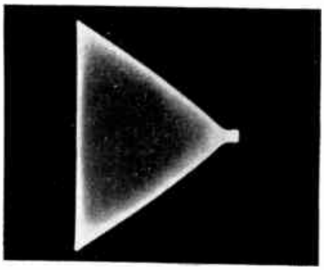
Modulation Monitoring

It is always desirable to modulate as fully as possible, but 100-per-cent modulation should not be exceeded — particularly in the downward direction — because harmonic distortion will be introduced and the channel width increased. This causes unnecessary interference to other stations. The oscilloscope is the best instrument for continuously checking the

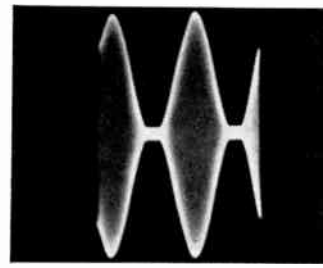


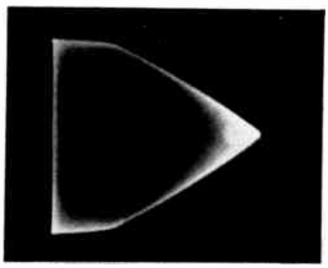
Properly-operated 'phone transmitter modulated 100 per cent.



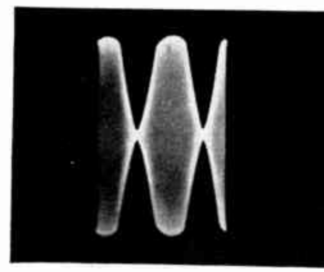


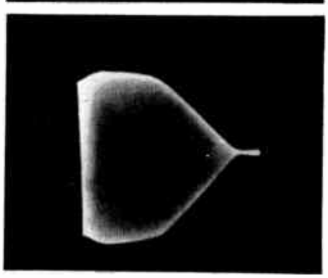
Overmodulation of a transmitter having high modulation capability. Distortion occurs only on the down-peaks.



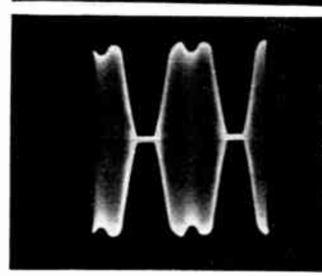


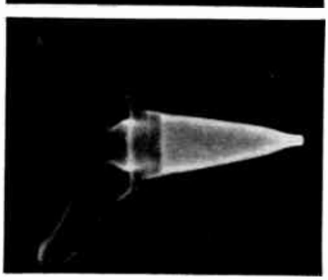
Nonlinearity in modulated r.f. stage, frequently caused by insufficient excitation of a plate-modulated amplifier or overexcitation of a grid-bias modulated amplifier. The amplifier modulates linearly in the downward direction but the up-peaks are flattened.



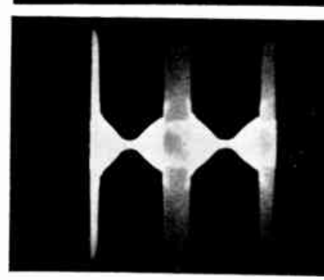


Overmodulation and nonlinear operation (insufficient modulation capability). These patterns are similar to those directly above, but with the modulation carried beyond 100 per cent in the downward direction.

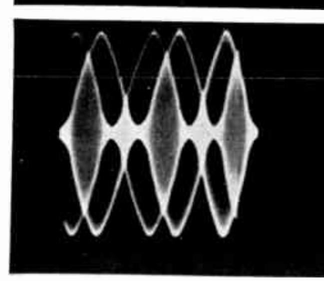




Overmodulation and parasitic oscillations in the modulated amplifier. The trapezoidal pattern also shows phase distortion caused by incorrect coupling between the oscilloscope and audio system.



Left — Phase distortion caused by incorrect coupling between audio system and oscilloscope. *Right* — Multiple pattern caused by incorrect setting of oscilloscope time-base control. In both cases the wave is modulated 100 per cent.



PHOTOGRAPHS OF TYPICAL OSCILLOSCOPE PATTERNS

These photographs show various conditions of modulation as displayed by the wedge or trapezoidal patterns in the left-hand column and the wave-envelope patterns in the right-hand column.

(Photographs reproduced through courtesy of the Allen B. DuMont Laboratories, Inc., Passaic, N. J.)

modulation. However, simpler indicators may be used for the purpose, once calibrated.

A convenient indicator, when a Class B modulator is used, is the plate milliammeter in the Class B stage, since plate current fluctuates with the voice intensity. Using the oscilloscope, determine the gain-control setting and voice intensity that give 100-per-cent modulation on voice peaks, and simultaneously observe the maximum Class B plate-milliammeter reading on the peaks. When this maximum reading is obtained, it will suffice to adjust the gain so that it is not exceeded.

A sensitive rectifier-type voltmeter (copper-oxide type) also can be used for modulation monitoring. It should be connected across the output circuit of an audio driver stage where the power level is a few watts, and similarly calibrated against the oscilloscope to determine the reading that represents 100-per-cent modulation.

The plate milliammeter of the modulated r.f. stage also is of some value as an indicator of overmodulation. The average plate current stays constant if the amplifier is linear, so the reading will be the same whether or not the transmitter is modulated. When the amplifier is overmodulated, especially in the downward direction, the operation is no longer linear and the average plate current will change. A flicker of the pointer may therefore be taken as an indication of overmodulation or nonlinearity. However, it is possible that under some operating conditions the average plate current will remain constant even though the amplifier is considerably overmodulated. Therefore an indicator of this type is not wholly reliable unless it has been checked previously against an oscilloscope.

Linearity

The linearity of a modulated amplifier may readily be checked with the oscilloscope. The trapezoidal pattern is more easily interpreted than the wave-envelope pattern, and less auxiliary equipment is required. The connections are the same as for measuring modulation percentage (Fig. 9-45B). If the amplifier is perfectly linear, the sloping sides of the trapezoid will be perfectly straight from the point at the axis up to at least 100-per-cent modulation in the upward direction. Nonlinearity will be shown by curvature of the sides. Curvature near the point, causing it to approach the axis more slowly than would occur with straight sides, indicates that the output power does not decrease rapidly enough in this region; it may also be caused by positive feed-back (a push-pull amplifier is recommended because better neutralization is possible than with single-ended amplifiers) or by r.f. leakage from the exciter through the final stage. The latter condition can be checked by removing the plate voltage from the modulated stage, when the carrier should disappear, leaving only the beam spot remaining on the screen (Fig. 9-46F). If a small vertical line remains, the amplifier should be reneutralized; if this does not eliminate the line, it is an indication that r.f. is being picked up from lower-power stages, either by coupling through the final tank or via the oscilloscope pick-up loop.

Inward curvature at the large end of the pattern is caused by improper operating conditions of the modulated amplifier — usually improper bias or insufficient excitation, or both, with plate modulation. In grid-bias and cathode-modulated systems, the bias, excitation and plate loading are not correctly proportioned when such curvature occurs. The usual reason is that the amplifier has been adjusted to have too-high carrier efficiency without modulation.

Fig. 9-47 shows typical patterns of both the trapezoid and wave-envelope types. The cause of the distortion is indicated for grid-bias and suppressor modulation. The patterns at A, although not truly linear, are representative of properly-operated grid-bias modulation systems. Better linearity can be obtained with plate modulation of a Class C amplifier.

Faulty Patterns

The drawings of Figs. 9-46 and 9-47 show what is normally to be expected in the way of pattern shapes when the oscilloscope is used to check modulation. If the actual patterns differ considerably from those shown, it may be that the pattern is faulty rather than the transmitter. It is important that only r.f. from the modulated stage be coupled to the oscilloscope, and then only to the vertical plates. The effect of stray r.f. from other stages in the transmitter has been mentioned in the preceding section. If r.f. is present also on the horizontal plates,

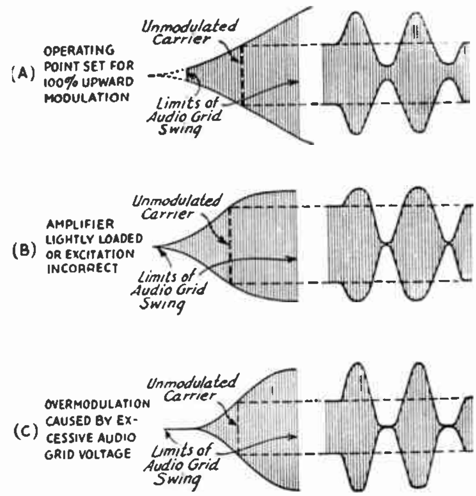


Fig. 9-47 — Oscilloscope patterns representing proper and improper adjustments for grid-bias or cathode modulation. Trapezoidal pattern at left; wave-envelope pattern at right. The pattern obtained with a correctly-adjusted amplifier is shown at A. The other drawings indicate nonlinear modulation from typical causes.

Spurious Sidebands

A superheterodyne receiver having a crystal filter is needed for checking spurious sidebands outside the normal communication channel. The r.f. input to the receiver must be kept low enough, by removing the antenna or by adequate separation from the transmitter, to avoid overloading and consequent spurious receiver responses. With the crystal filter in its sharpest position and the beat oscillator turned on, tune through the region outside the normal channel limits (3 to 4 kilocycles each side of the carrier) while another person talks into the microphone. Spurious sidebands will be observed as intermittent beat notes coinciding with voice peaks — or, in bad cases of distortion or overmodulation, as "clicks" or crackles well away from the carrier frequency. Sidebands more than 3 to 4 kilocycles from the carrier should be of negligible strength in a properly-modulated 'phone transmitter. The causes are overmodulation or nonlinear operation.

R.F. in Speech Amplifier

A small amount of r.f. current in the speech amplifier — particularly in the first stage, which is most susceptible to such r.f. pick-up — will cause overloading and distortion in the low-level stages. Frequently also there is a regenerative effect which causes an audio-frequency oscillation or "howl" to be set up in the audio system. In such cases the gain control cannot be advanced very far before the howl builds up, even though the amplifier may be perfectly stable when the r.f. section of the transmitter is not turned on.

Complete shielding of the microphone, microphone cord, and speech amplifier is necessary to prevent r.f. pick-up, and a ground connection separate from that to which the transmitter is connected is advisable. Direct coupling or unsymmetrical coupling to the antenna (single-wire feed, feeders tapped on final tank circuit, etc.) may be responsible because these systems sometimes cause the transmitter chassis to take an r.f. potential above ground. Inductive coupling to a two-wire transmission line is advisable. This antenna effect can be checked by disconnecting the antenna and dissipating the r.f. power in a dummy antenna, when it usually will be found that the r.f. feed-back disappears. If it does not, the speech amplifier and microphone shielding are at fault.

Overmodulation Indicators

The most positive method of *preventing* overmodulation is the clipper-filter system described earlier, when properly set up and adjusted. In the absence of such a system — or even with it, just to be safe — some form of overmodulation indicator should be in constant use when the transmitter is on the air.

The best device for this purpose is the cathode-ray oscilloscope. The trapezoidal and wave-envelope patterns are equally useful. A 60-cycle sinusoidal sweep will be quite satisfactory for the wave-envelope pattern. Either pattern should be watched particularly for the bright spots at the axis that accompany overmodulation in the downward direction. The speaking-voice intensity should be kept below the level that shows 100-per-cent modulation on the 'scope.

Overmodulation on negative peaks is more likely to result in spurious sidebands than overmodulation in the upward direction because of the sharp break that occurs when the carrier is suddenly cut off and on. The milliammeter in the negative-peak indicator of Fig. 9-48 will show a reading on each overmodulation peak that carries the instantaneous voltage on the plate of the Class C modulated amplifier "below zero" — that is, negative. The rectifier, *V*, cannot conduct so long as the

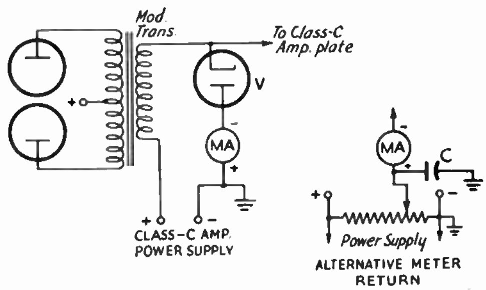


Fig. 9-48 — A negative-peak overmodulation indicator. Milliammeter *M.1* may be any low-range instrument (up to 0–50 ma. or so). The inverse-peak-voltage rating of the rectifier, *V*, must be at least equal to the d.c. voltage applied to the plate of the r.f. amplifier. The alternative meter-return circuit can be used to indicate modulation in excess of any desired value below 100 per cent. The reactance of the by-pass condenser, *C*, at 100 cycles should be small compared with the resistance across which it is connected. An 8-μfd. electrolytic condenser will be satisfactory if the resistance it shunts is 1000 ohms or more.

negative half-cycle of audio output voltage is less than the d.c. voltage applied to the r.f. tube.

The inverse-peak-voltage rating of the rectifier tube must be at least twice the d.c. voltage applied to the plate of the modulated Class C amplifier. The filament transformer likewise must have insulation rated to withstand twice the d.c. plate voltage. Either mercury-vapor or high-vacuum rectifiers can be used. The 15-volt breakdown voltage of the former will introduce a slight error, since the plate voltage must go at least 15 volts negative before the rectifier will ionize, but the error is inconsequential at plate voltages above a few hundred volts.

The effectiveness of the monitor is improved if it indicates at somewhat less than 100-per-cent modulation, as it will then warn of the danger of overmodulation before it actually occurs. It can be adjusted to indicate at any the pattern will lean to one side instead of being upright. If the oscilloscope cannot be moved to a spot where the unwanted pick-up disappears, a small by-pass condenser (10 $\mu\mu$fd.) should be connected across the horizontal plates as close to the cathode-ray tube as possible. An r.f. choke (2.5 mh. or smaller) may also be connected in series with the ungrounded horizontal plate.

"Folded" trapezoidal patterns, and patterns in which the sides of the trapezoid are elliptical instead of straight, occur when the audio sweep voltage is taken from some point in the audio system other than that where the a.f. power is applied to the modulated stage. Such patterns are caused by a phase difference between the sweep voltage and the modulating voltage. The connections should always be as shown in Fig. 9-45B.

Plate-Current Shift

As mentioned above, the d.c. plate current of a modulated amplifier will be the same with and without modulation so long as the amplifier operation is perfectly linear and other conditions remain unchanged. This also assumes that the modulator is working within its capabilities. Because there is usually some curvature of the modulation characteristic with grid-bias modulation there is normally a slight upward change in plate current of a stage so modulated, but this occurs only at high modulation percentages and is barely detectable under the usual conditions of voice modulation.

With plate modulation, a downward shift in plate current may indicate one or more of the following:

1) Insufficient excitation to the modulated r.f. amplifier.
2) Insufficient grid bias on the modulated stage.
3) Wrong load resistance for the Class C r.f. amplifier.
4) Insufficient output capacitance in the filter of the modulated-amplifier plate supply.
5) Heavy overloading of the Class C r.f. amplifier tube or tubes.

Any of the following may cause an upward shift in plate current:

1) Overmodulation (excessive audio power, audio gain too great).
2) Incomplete neutralization of the modulated amplifier.
3) Parasitic oscillation in the modulated amplifier.

When a common plate supply is used for both a Class B (or Class AB) modulator and a modulated r.f. amplifier, the plate current of the latter may "kick" downward because of poor power-supply voltage regulation with the varying additional load of the modulator on the supply. The same effect may occur with high-power transmitters because of poor regulation of the a.c. supply mains, even when a separate power-supply unit is used for the Class B modulator. Either condition may be detected by measuring the plate voltage applied to the modulated stage; in addition, poor line regulation also may be detected by observing if there is any downward shift in filament or line voltage.

With grid-bias modulation, any of the following may be the cause of a plate-current shift greater than the normal mentioned above:

Downward kick: Too much r.f. excitation; insufficient operating bias; distortion in modulator or speech amplifier; too-high resistance in bias supply; insufficient output capacitance in plate-supply filter to modulated amplifier; amplifier plate circuit not loaded heavily enough; plate-circuit efficiency too high under carrier conditions.

Upward kick: Overmodulation (excessive audio voltage); distortion in audio system; regeneration because of incomplete neutralization; operating grid bias too high.

A downward kick in plate current will accompany an oscilloscope pattern like that of Fig. 9-47B; the pattern with an upward kick will look like Fig. 9-47A, with the shaded portion extending farther to the right and above the carrier, for the "wedge" pattern.

Noise and Hum on Carrier

Noise and hum may be detected by listening to the signal on a receiver, provided the receiver is far enough away from the transmitter to avoid overloading. The hum level should be low compared to the voice at 100-per-cent modulation. Hum may come either from the speech amplifier and modulator or from the r.f. section of the transmitter. Hum from the r.f. section can be detected by completely shutting off the modulator; if hum remains when this is done, the power-supply filters for one or more of the r.f. stages have insufficient smoothing. With a hum-free carrier, hum introduced by the modulator can be checked by turning on the modulator but leaving the speech amplifier off; power-supply filtering is the likely source of such hum. If carrier and modulator are both clean, connect the speech amplifier and observe the increase in hum level. If the hum disappears with the gain control at minimum, the hum is being introduced in the stage or stages preceding the gain control. The microphone also may pick up hum, a condition that can be checked by removing the microphone from the circuit but leaving the first speech-amplifier grid circuit otherwise unchanged. A good ground on the microphone and speech system usually is essential to hum-free operation.

Hum can be checked with the oscilloscope, where it has the same appearance as ordinary modulation on the carrier. While the percentage usually is rather small, if the carrier shows modulation with no speech input hum is the likely cause. The various parts of the transmitter may be checked through as described above.

desired modulation percentage by making the meter return to a point on the power-supply bleeder as shown in the alternative diagram. The by-pass condenser, *C*, insures that the full audio voltage appears across the indicator circuit. The modulation percentage at which the system indicates is determined by the ratio of the d.c. voltage between the milliammeter tap and the positive terminal to the total d.c. voltage.

Frequency and Phase Modulation

The primary advantage of frequency modulation (FM) or phase modulation (PM) over amplitude modulation (AM) comes from the fact that noise or "static," whether natural or set up by electrical machines, is fundamentally an amplitude effect. An AM detector responds to noise just as readily as to the desired modulation on a signal. However, if the receiving system responds principally to frequency or phase changes and is insensitive to amplitude variations, it will give normal reception of an FM or PM signal but noise will be greatly reduced.

The improvement that can be realized by using FM or PM instead of AM depends on the strength of the received signal, the character of the noise, and the way the noise is distributed over the receiver passband. In general, the wider the channel occupied by the signal the better the noise suppression — if the signal strength is above a certain **threshold** value. The wider the channel occupied by the signal, the stronger the signal required to reach the threshold.

In amateur work, FM and PM have been used not so much because of the possibility of an improved signal-to-noise ratio (the improvement is not great with narrow bandwidths) but because of more-or-less incidental advantages. For example, in the ultrahigh and superhigh frequency ranges some tubes do not lend themselves well to amplitude modulation, but can easily be frequency-modulated. On the lower frequencies FM and PM are often used because they cause less interference than AM in unshielded broadcast receivers in the vicinity.

Frequency Modulation

Fig. 9-49 is a representation of frequency modulation. When a modulating signal is applied, the carrier frequency is increased during one half-cycle of the modulating signal and decreased during the half-cycle of opposite polarity. This is indicated in the drawing by the fact that the r.f. cycles occupy less time (higher frequency) when the modulating signal is positive, and more time (lower frequency) when the modulating signal is negative. The change in the carrier frequency **(frequency deviation)** is proportional to the instantaneous amplitude of the modulating signal, so the deviation is small when the instantaneous amplitude of the modulating signal is small, and is greatest when the modulating signal reaches its peak, either positive or negative. That is, the frequency deviation follows the instantaneous changes in the amplitude of the modulating signal.

As shown by the drawing, the amplitude of the signal does not change during modulation.

Phase and Frequency

To understand the difference between FM and PM it is necessary to appreciate that the frequency of an alternating current is determined by the *rate at which its phase changes.* A current in which the phase changes rapidly has a higher frequency than one in which the phase changes slowly. For example, if the phase moves through 360 degrees in one second the frequency is one cycle per second, but if the phase moves through 1080 degrees in one second (3×360 degrees) there are three complete cycles in one second.

If the phase of the current in a circuit is changed — this might be done by adjusting the tuning of an amplifier tank circuit, for example — there is an instantaneous frequency change during the time that the phase is being shifted. The amount of frequency change, or deviation, depends on how rapidly the phase shift is accomplished. It is also dependent upon the total amount of the phase shift. In a properly-operating PM system the amount of phase shift is proportional to the instantaneous amplitude of the modulating signal. The rapidity of the phase shift is directly proportional to the frequency of the modulating signal. Consequently, the frequency deviation in PM is proportional to both the amplitude and frequency of the modulating signal. The latter

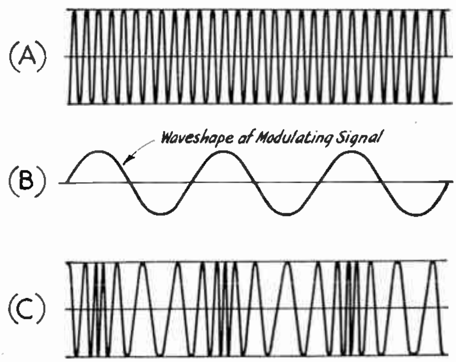


Fig. 9-49 — Graphical representation of frequency modulation. In the unmodulated carrier at A, each r.f. cycle occupies the same amount of time. When the modulating signal, B, is applied, the radio frequency is increased and decreased according to the amplitude and polarity of the modulating signal.

represents the outstanding difference between FM and PM, since in FM the frequency deviation is proportional only to the amplitude of the modulating signal.

Modulation Depth

In FM or PM there is no condition that corresponds exactly to overmodulation in AM. "Percentage of modulation" has to be defined a little differently for these systems. Practically, "100-per-cent modulation" is reached when the transmitted signal occupies a channel just equal to the bandwidth for which the *receiver* is designed. If the channel occupied is wider than the receiver can accept, the receiver distorts the signal and the end effect is much the same as overmodulation in AM. However, on another receiver designed for a different bandwidth the same signal might be equivalent to only 25-per-cent modulation.

In amateur work no specifications have been set up for channel width except in the case of "narrow-band" FM or PM (frequently abbreviated NFM), where the channel width is defined as being the same as that of a properly-modulated AM signal. That is, the channel width for NFM does not exceed twice the highest audio frequency in the modulating signal. NFM transmissions based on an upper audio limit of 3000 cycles therefore should occupy a channel no wider than 6 kc.

FM and PM Sidebands

It might be surmised that the channel occupied by an FM or PM signal is no greater than the frequency deviation on both sides of the carrier. Similar reasoning applied to amplitude modulation would lead to the conclusion that an AM signal takes up no more space than the carrier alone, since only the *amplitude* of the carrier varies. However, the fact is that both FM and PM set up sidebands, just as AM does. In the case of FM and PM, single-tone modulation sets up a whole series of pairs of sidebands that are harmonically related to the modulating frequency, whereas in AM there is only one pair of sidebands.

The number of "extra" sidebands that occur in FM and PM depends on the relationship between the modulating frequency and the carrier frequency deviation. The ratio between the frequency deviation, in cycles per second, and the modulating frequency, also in cycles per second, is called the **modulation index.** That is,

$$Modulation\ index = \frac{Carrier\ frequency\ deviation}{Modulating\ frequency}$$

Example: The maximum frequency deviation in an FM transmitter is 3000 cycles either side of the carrier frequency. The modulation index when the modulating frequency is 1000 cycles is

$$\text{Modulation index} = \frac{3000}{1000} = 3$$

At the same deviation with 3000-cycle modulation the index would be 1; at 100 cycles it would be 30, and so on.

The modulation index is also equal to the phase shift in radians. In PM the index is constant regardless of the modulating frequency; in FM it varies with the modulating frequency, as shown in the previous example. To identify any particular FM system, the limiting modulation index — that is, the ratio of the *maximum* carrier-frequency deviation to the *highest* modulating frequency used — is called the **deviation ratio.**

Fig. 9-50 shows how the amplitudes of the carrier and the various sidebands vary with the modulation index. This is for single-tone modulation; the first sideband (actually a pair, one above and one below the carrier) is displaced from the carrier by an amount equal to the modulating frequency, the second is twice the modulating frequency away from the carrier, and so on. For example, if the modulating frequency is 2000 cycles and the carrier frequency is 29,500 kc., the first sideband pair is at 29,498 kc. and 29,502 kc., the second pair is at 29,496 kc. and 29,504 kc., the third at 29,494 kc. and 29,506 kc., etc. The amplitudes of these sidebands depend on the modulation index, not on the frequency deviation. In AM, regardless of the percentage of modulation (so long as it does not exceed 100 per cent) the sidebands would appear *only* at 29,498 and 29,502 kc. under the same conditions.

Note that, as shown by Fig. 9-50, the carrier strength varies with the modulation index. (In amplitude modulation the carrier strength is constant; only the sideband amplitude varies.) At a modulation index of approximately 2.4 the carrier disappears entirely and then becomes "negative" at a higher index. This simply means that its phase is reversed as compared to the phase without modulation. In FM and PM the energy that goes into the sidebands is taken from the carrier, the *total* power remaining the same regardless of the modulation index. In AM the sideband power is supplied by the modulator in the case of

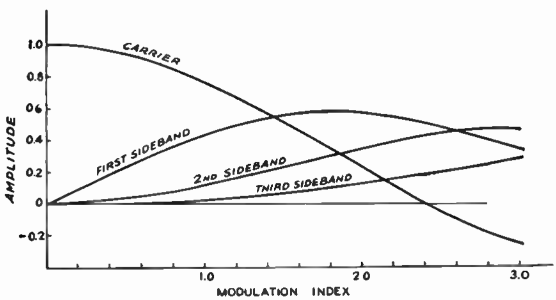


Fig. 9-50 — How the amplitude of the pairs of sidebands varies with the modulation index in an FM or PM signal. If the curves were extended for greater values of modulation index it would be seen that the carrier amplitude goes through zero at several points. The same statement also applies to the sidebands.

plate modulation, and by changing the power input and efficiency in the case of grid-bias modulation.

The curves of Fig. 9-50 can be carried out to considerably-higher modulation indexes, in which case it will be discovered that more and more additional sidebands are set up and that the carrier goes through several "zeros" and reversals in phase.

Frequency Multiplication

In frequency or phase modulation there is no change in the amplitude of the signal with modulation, consequently an FM or PM signal can be amplified by an ordinary Class C amplifier without distortion. The modulation can take place in a very low-level stage and the signal can then be amplified by either frequency multipliers or straight amplifiers. The audio power required for modulating an FM or PM transmitter is negligible.

If the modulated signal is passed through one or more frequency multipliers, the modulation index is multiplied by the same factor that the carrier frequency is multiplied. For example, suppose that modulation is applied on 3.5 Mc. and the final output is on 28 Mc. The total frequency multiplication is 8 times, so if the frequency deviation is 500 cycles at 3.5 Mc., it will be 4000 cycles at 28 Mc. Frequency multiplication offers a means for obtaining practically any desired amount of frequency deviation, whether or not the modulator itself is capable of giving that much deviation without distortion.

Where FM or PM is used in crowded 'phone bands (particularly below 29 Mc.) it is of utmost importance that the transmissions should occupy a channel no wider than would be occupied by an AM signal. It is evident from Fig. 9-50 that this requirement can be met only by using a relatively small modulation index. It must be realized that the higher-order sidebands always are present, even at very small indexes. If the modulation index (with single-tone modulation) does not exceed about 0.6 the most important extra sideband, the second, will be at least 20 db. below the unmodulated carrier level, and this should represent an effective channel width about equivalent to that of an AM signal. In the case of speech, a somewhat higher modulation index can be used. This is because the energy distribution in a complex wave is such that the modulation index for any one frequency component is reduced, as compared to the index with a sine wave having the same peak amplitude as the voice wave.

The chief advantage of narrow-band FM or PM for frequencies below 30 Mc. is that it eliminates or reduces certain types of interference to broadcast reception. Also, the modulating equipment is relatively simple and inexpensive. However, assuming the same unmodulated carrier power in all cases, narrow-band FM or PM is not as effective as AM. As shown by Fig. 9-50, at an index of 0.6 the amplitude of the first sideband is about 25 per cent of the unmodulated-carrier amplitude; this compares with a sideband amplitude of 50 per cent in the case of a 100-per-cent modulated AM transmitter. In other words, so far as effectiveness is concerned, a narrow-band FM or PM transmitter is about equivalent to a 100-per-cent modulated AM transmitter operating at one-fourth the power input.

Comparison of FM and PM

The methods used by amateurs for the reception of FM or PM signals (see Chapter Five) are for the most part better adapted to frequency modulation than to phase modulation. On a receiver properly adjusted for FM reception the outstanding difference between the two systems is that FM sounds natural, while a PM signal lacks "lows." This is because, for a given receiver bandwidth, the audio output from a receiver set for FM reception is proportional to the frequency deviation. In FM transmission the deviation is the same for all audio frequencies of the same amplitude, but in PM the deviation is proportional to the audio frequency. Hence if a 3000-cycle modulating signal of given amplitude results in a certain frequency deviation, a 100-cycle modulating signal of the same amplitude will give only one-thirtieth as much deviation. The crystal-filter receiving method described in Chapter Five overcomes this, but is not used by many amateurs because the adjustment is somewhat critical.

Frequency modulation cannot be applied to an amplifier stage, but phase modulation can. PM is therefore readily adaptable to transmitters employing oscillators of high stability such as the crystal-controlled type. The amount of phase shift that can be obtained with good linearity is limited to about one-half radian; in other words, the maximum practicable modulation index is 0.5 at the radio frequency at which the modulation takes place. Because the phase shift is proportional to the modulating frequency, this index can be used only at the highest frequency present in the modulating signal, assuming that all frequencies will at one time or another have equal amplitudes. Taking 3000 cycles as a suitable upper limit for voice work, and setting the modulation index at 0.5 for 3000 cycles, the frequency response of the speech-amplifier system above 3000 cycles must be sharply attenuated, to prevent sideband splatter. Also, if the "tinny" quality of PM as received on an FM receiver is to be avoided, the PM must be changed to FM, in which the modulation index decreases in inverse proportion to the modulating frequency. This requires shaping the speech-amplifier frequency-response curve in such a way that the output voltage is inversely proportional to frequency, at least over the voice range. When this is done the maximum modulation index can only be used at the lowest audio frequency, approximately 100 cycles in voice transmission, and must decrease in proportion to the increase in frequency. The result is that the maximum linear frequency deviation is only about 50 cycles, when PM is changed to FM. To increase the deviation to 3000 cycles requires a frequency multiplication of 3000/50, or 60 times.

In contrast, it is relatively easy to secure a fairly-large frequency deviation when a self-controlled oscillator is frequency-modulated directly. (True frequency modulation of a crystal-controlled oscillator results in only very small deviations and so requires a great deal of frequency multiplication.) The chief problem is to maintain a satisfactory degree of carrier stability, since the greater the inherent stability of the oscillator the more difficult it is to secure a wide frequency swing with linearity. However, it is possible, with a compromise design, to secure a frequency deviation of 3000 cycles at all amateur frequencies on which FM is permitted. It is very easy to do so at 14 Mc. and higher, especially when the oscillator frequency is such that a frequency multiplication of 4 or more is possible.

Methods of Frequency and Phase Modulation

FREQUENCY MODULATION

The simplest and most satisfactory device for amateur FM is the reactance modulator. This is a vacuum tube connected to the r.f. tank circuit of an oscillator in such a way as to act as a variable inductance or capacitance. Fig. 9-51 is a representative circuit. The control-grid circuit of the 6L7 tube is connected across the small capacitance, C_1, which is in series with the resistor, R_1, across the oscillator tank circuit. Any type of oscillator circuit may be used. The resistance of R_1 is made large compared to the reactance of C_1, so the r.f. current through R_1C_1 will be practically in phase with the r.f. voltage appearing at the terminals of the tank circuit. However, the voltage across C_1 will lag the current by 90 degrees. The r.f. current in the plate circuit of the 6L7 will be in phase with the grid voltage, and consequently is 90 degrees behind the current through C_1, or 90 degrees behind the r.f. tank voltage. This lagging current is drawn through the oscillator tank, giving the same effect as though an inductance were connected across the tank. The frequency increases in proportion to the amplitude of the lagging plate current of the modulator. The value of plate current is determined by the voltage on the No. 3 grid of the 6L7; hence the oscillator frequency will vary when an audio signal voltage is applied to the No. 3 grid.

If, on the other hand, C_1 and R_1 are interchanged and the reactance of C_1 is made large compared to the resistance of R_1, the r.f. current in the 6L7 plate circuit will lead the oscillator-tank r.f. voltage, making the reactance capacitive rather than inductive.

A circuit using a receiving-type r.f. pentode of the high-transconductance type, such as the 6SG7, is shown in Fig. 9-52. In this case, both r.f. and audio are applied to the control grid. The audio voltage, introduced through a radio-frequency choke, *RFC*, varies the transconductance of the tube and thereby varies the r.f. plate current. The capacitance C_8 corresponds to C_1 in Fig. 9-51; it represents the input capacitance of the tube. (It is possible, also, to omit C_1 from Fig. 9-51 and depend upon the input capacitance of the 6L7 instead; the only disadvantage is that there is then no control over the modulator sensitivity. Likewise, a 3-30-$\mu\mu$fd. trimmer condenser can be connected at C_8 in Fig. 9-52 to permit controlling the sensitivity.) In Fig. 9-52 the r.f. circuit is series-fed, which is advantageous if the r.f. tube and the modulator can be operated at the same plate voltage. The use of different plate voltages on the two tubes calls for the parallel-feed arrangement shown in Fig. 9-51.

The modulated oscillator usually is operated on a relatively low frequency, so that a high order of carrier stability can be secured. Frequency multipliers are used to raise the frequency to the final frequency desired. The frequency deviation increases with the number of times the initial frequency is multiplied; for instance, if the oscillator is operated on 6.5 Mc. and the output frequency is to be 52 Mc., an oscillator frequency deviation of 1000 cycles will be raised to 8000 cycles at the output frequency.

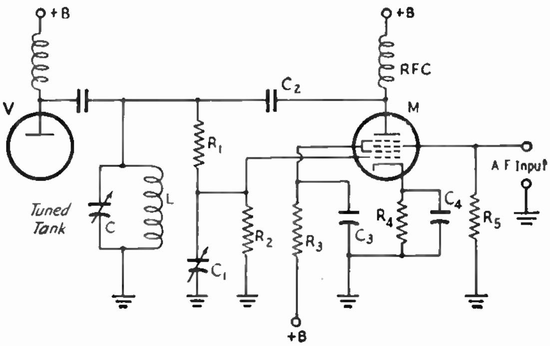


Fig. 9-51 — Reactance-modulator circuit using a 6L7 tube.

C — R.f. tank capacitance. C_1 — 3-30 $\mu\mu$fd. C_2 — 220 $\mu\mu$fd.
C_3 — 8-μfd. electrolytic (a.f. by-pass) in parallel with 0.01-μfd. paper (r.f. by-pass).
C_4 — 10-μfd. electrolytic in parallel with 0.01-μfd. paper.
L — R.f. tank inductance.
R_1 — 47,000 ohms.
R_2, R_5 — 0.47 megohm.
R_3 — 33,000 ohms.
R_4 — 330 ohms.
RFC — 2.5 mh.

A reactance modulator can be connected to a crystal oscillator as well as to the self-controlled type. However, the resulting signal is more phase-modulated than it is frequency-modulated, for the reason that the frequency deviation that can be secured by varying the tuning of a crystal oscillator is quite small.

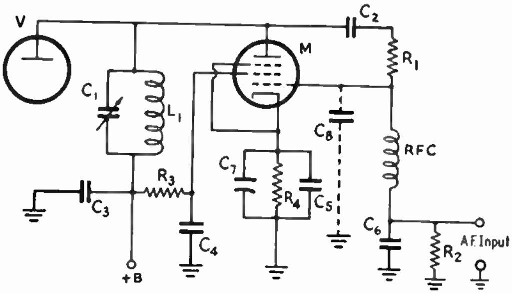


Fig. 9-52 — Reactance modulator using a high-transconductance pentode (6SG7, 6AG7, etc.).

C_1 — R.f. tank capacitance (see text).
C_2, C_3 — 0.001-μfd. mica.
C_4, C_5, C_6 — 0.0047-μfd. mica.
C_7 — 10-μfd. electrolytic.
C_8 — Tube input capacitance (see text).
R_1, R_2 — 0.47 megohm.
R_3 — Screen dropping resistor; select to give proper screen voltage on type of modulator tube used.
R_4 — Cathode bias resistor; select as in case of R_3.
L_1 — R.f. tank inductance.
RFC — 2.5-mh. r.f. choke.

Design Considerations

The sensitivity of the modulator (frequency change per unit change in grid voltage) depends on the transconductance of the modulator tube. It increases when C_1 is made smaller, for a fixed value of R_1, and also increases with an increase in L/C ratio in the oscillator tank circuit. Since the carrier stability of the oscillator depends on the L/C ratio, it is desirable to use the highest tank capacitance that will permit the desired deviation to be secured while keeping within the limits of linear operation. When the circuit of Fig. 9-52 is used in connection with a 7-Mc. oscillator, a linear deviation of 1500 cycles above and below the carrier frequency can be secured when the oscillator tank capacitance is approximately 200 μμfd. A peak a.f. input of two volts is required for full deviation.

A change in *any* of the voltages on the modulator tube will cause a change in r.f. plate current, and consequently a frequency change. Therefore it is advisable to use a regulated plate power supply for both modulator and oscillator. At the low voltages used (250 volts) the required stabilization can be secured by means of gaseous regulator tubes.

Speech Amplification

The speech amplifier preceding the modulator follows ordinary design, except that no power is required from it and the a.f. voltage taken by the modulator grid usually is small — not more than 10 or 15 volts, even with large modulator tubes. Because of these modest requirements, only a few speech-amplifier stages are needed; a two-stage amplifier consisting of a pentode followed by a triode, both resistance-coupled, will more than suffice for crystal microphones.

PHASE MODULATION

The same type of reactance-tube circuit that is used to vary the tuning of the oscillator tank in FM can be used to vary the tuning of an amplifier tank and thus vary the phase of the tank current for PM. Hence the modulator circuits of Figs. 9-51 and 9-52 can be used for PM if the reactance tube works on an amplifier tank instead of directly on a self-controlled oscillator.

The phase shift that occurs when a circuit is detuned from resonance depends on the amount of detuning and the Q of the circuit. The higher the Q, the smaller the amount of detuning needed to secure a given number of degrees of phase shift. If the Q is at least 10, the relationship between phase shift and detuning (in kilocycles either side of the resonant frequency) will be substantially linear over a range of about 25 degrees. From the standpoint of modulator sensitivity, the Q of the tuned circuit on which the modulator operates should be as high as possible. On the other hand, the effective Q of the circuit will not be very high if the amplifier is delivering power to a load since the load resistance reduces the Q. There must therefore be a compromise between modulator sensitivity and r.f. power output from the modulated amplifier. An optimum figure for Q appears to be about 20; this allows reasonable loading of the modulated amplifier and the necessary tuning variation can be secured from a reactance modulator without difficulty. It is advisable to modulate at a very low power level — preferably in a transmitter stage where receiving-type tubes are used.

Reactance modulation of an amplifier stage usually also results in simultaneous amplitude modulation. This must be eliminated by feeding the modulated signal through an amplitude limiter or one or more "saturating" stages — that is, amplifiers that are operated Class C and driven hard enough so that variations in the amplitude of the grid excitation produce no appreciable variations in the final output amplitude.

For the same type of reactance modulator, the speech-amplifier gain required is the same for PM as for FM. However, as pointed out earlier, the fact that the actual frequency deviation increases with the modulating audio frequency in PM makes it necessary to cut off the frequencies above about 3000 cycles before modulation takes place. If this is not done, unnecessary sidebands will be generated at frequencies considerably away from the carrier.

Reactance-Modulator Unit for Narrow-Band FM

The FM speech-amplifier and modulator unit shown in Figs. 9-53 and 9-54 uses a pentode reactance modulator in a circuit which is basically that of Fig. 9-52. It differs only in the detail that the audio signal is applied to the control grid in parallel with the r.f. voltage from the oscillator, instead of the series-feed arrangement shown in Fig. 9-52. Because of the parallel feed, resistor R_4 is incorporated in the circuit to prevent r.f. from appearing in the plate circuit of the speech-amplifier tube.

The unit uses miniature tubes for the sake of making a compact assembly that can be mounted in any convenient spot near the VFO tuned circuit. In Fig. 9-53 it is shown mounted on the outside of the VFO case. When this type of mounting is used the unit should be placed so that the lead between the VFO tuned circuit and the modulator is as short as possible. If there is space available, it is preferable to mount the unit inside the VFO cabinet.

The chassis for the unit is 4 inches long by 2 inches wide, and has a mounting lip 2 inches deep. As shown in the photographs, it is formed from a piece of aluminum with the edges turned over to stiffen it. The various components are easily accommodated underneath. The r.f. leads should be kept short and separated as much as possible from the audio and power-supply wiring.

Filament and plate power can usually be taken from the VFO supply, since the total plate current is only a few milliamperes. Filament current required is 0.6 amp. The microphone input is carried through a shielded lead to the unit, thus the microphone connector can be placed in any convenient location on the VFO unit itself. Once the proper setting of the gain control is found it need not be touched again, so screwdriver adjustment is quite adequate.

The adjustment of reactance modulators is discussed in a later section in this chapter.

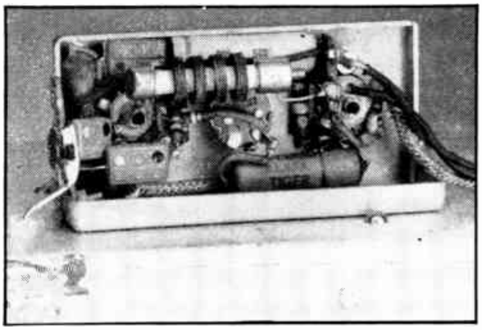


Fig. 9-54 — Underneath the modulator unit. The r.f. connection to the VFO goes through the feed-through bushing at the left.

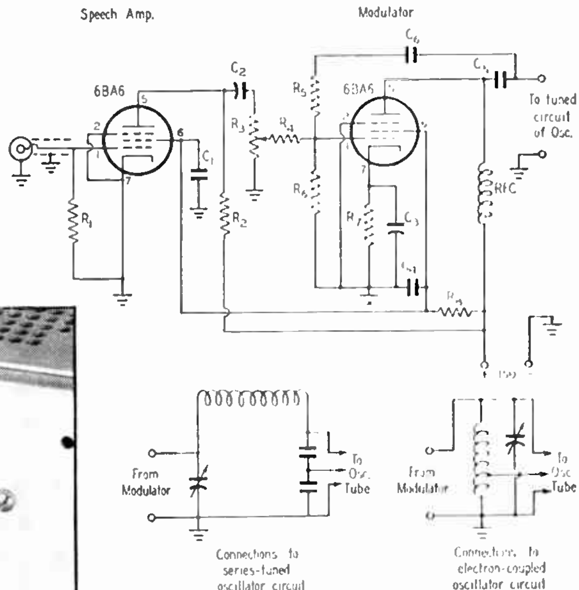


Fig. 9-55 — Circuit diagram of the narrow-band FM modulator unit.

C_1 — 680-μμfd. mica.
C_2, C_4 — 0.01-μfd. paper, 100 volts.
C_3 — 0.025-μfd. paper, 200 volts.
C_5, C_6 — 47-μμfd. mica.
R_1 — 1.2 megohms, ½ watt.
R_2, R_8 — 0.22 megohm, ½ watt.
R_3 — 0.5-megohm potentiometer.
R_4 — 0.1 megohm, ½ watt.
R_5 — 10,000 ohms, ½ watt.
R_6 — 0.47 megohm, ½ watt.
R_7 — 390 ohms, ½ watt.
RFC — 2.5-mh. r.f. choke.

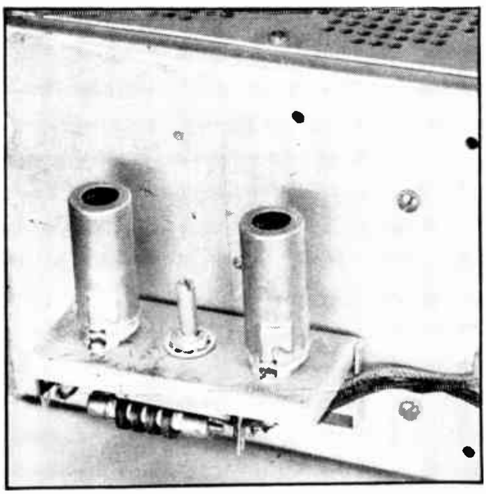


Fig. 9-53 — Miniature reactance modulator that can be used with any VFO. The shielded lead is for microphone input; the other two wires bring in filament and plate supply.

Checking FM and PM Transmitters

Accurate checking of the operation of an FM or PM transmitter requires different methods than the corresponding checks on an AM set. This is because the common forms of measuring devices either indicate amplitude variations only (a d.c. milliammeter, for example), or because their indications are most easily interpreted in terms of amplitude. There is no simple instrument that indicates frequency deviation in a modulated signal directly.

However, there is one favorable feature in FM or PM checking. The modulation takes place at a very low level and the stages following the one that is modulated do not affect the linearity of modulation so long as they are properly tuned. Therefore the modulation may be checked *without putting the transmitter on the air*, or even on a dummy antenna. The power is simply cut off the amplifiers following the modulated stage. This not only avoids unnecessary interference to other stations during testing periods, but also keeps the signal at such a

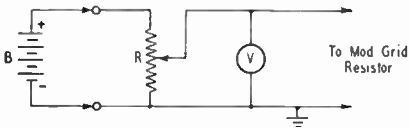


Fig. 9-56 — **D.c. method of checking frequency deviation of a reactance-tube-modulated oscillator. A 500- or 1000-ohm potentiometer may be used at *R*.**

low level that it may be observed quite easily on the station receiver. A good receiver with a crystal filter is an essential part of the checking equipment of an FM or PM transmitter, particularly for narrow-band FM or PM.

The quantities to be checked in an FM or PM transmitter are the linearity and frequency deviation. Because of the essential difference between FM and PM the methods of checking differ in detail.

Reactance-Tube FM

It was explained earlier that in FM the frequency deviation is the same at any audio modulation frequency if the audio signal amplitude does not vary. Since this is true at *any* audio frequency it is true at zero frequency. Consequently it is possible to calibrate a reactance modulator by applying an adjustable d.c. voltage to the modulator grid and noting the change in oscillator frequency as the voltage is varied. A suitable circuit for applying the adjustable voltage is shown in Fig. 9-56. The battery, *B*, should have a voltage of 3 to 6 volts (two or more dry cells in series). The arrows indicate clip connections so that the battery polarity can be reversed.

The oscillator frequency deviation should be measured by using a receiver in conjunction with an accurately-calibrated frequency meter, or by any means that will permit accurate measurement of frequency differences of a few hundred cycles. One simple method is to tune in the oscillator on the receiver (disconnecting the receiving antenna, if necessary, to keep the signal strength well below the overload point) and then set the receiver b.f.o. to zero beat. Then increase the d.c. voltage applied to the modulator grid from zero in steps of about ½ volt and note the beat frequency at each change. Then reverse the battery terminals and repeat. The frequency of the beat note may be measured by comparison with a calibrated audio-frequency oscillator, or by comparison with a piano or other musical instrument (see Chapter Twenty-Four for frequencies of musical tones). Note that with the battery polarity positive with respect to ground the radio frequency will move in one direction when the voltage is increased, and in the other direction when the battery terminals are reversed. When a number of readings has been taken a curve may be plotted to show the relationship between grid voltage and frequency deviation.

A sample curve is shown in Fig. 9-57. The usable portion of the curve is the center part which is essentially a straight line. The bending at the ends indicates that the modulator is no longer linear; this departure from linearity will cause harmonic distortion and will broaden the channel occupied by the signal. In the example, the characteristic is linear 1.5 kc. on either side of the center or carrier frequency. This is the maximum deviation permissible at the frequency at which the measurement is made. At the final output frequency the deviation will be multiplied by the same number of times that the measurement frequency is multiplied. This must be kept in mind when the check is made at a frequency that differs from the output frequency.

A good modulation indicator is a "magic-eye" tube such as the 6E5. This should be connected across the grid resistor of the reactance modulator as shown in Fig. 9-58. Note its deflection (using the d.c. voltage method as in

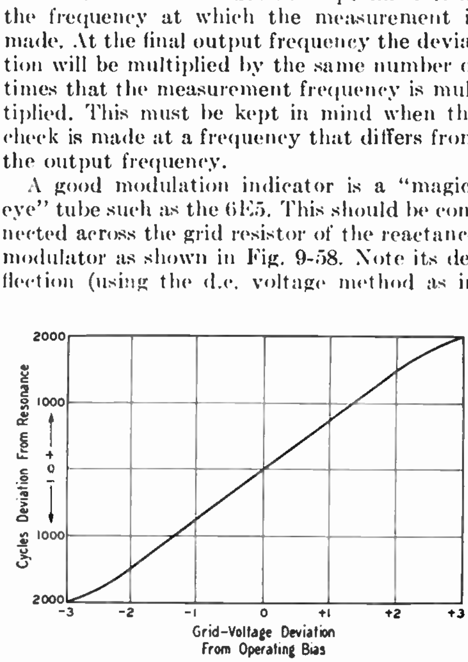


Fig. 9-57 — **A typical curve of frequency deviation *vs.* modulator grid voltage.**

Fig. 9-56) at the maximum deviation to be used. This deflection represents "100-per-cent modulation" and with speech input the gain should be kept at the point where it is just reached on voice peaks. If the transmitter is used on more than one band, the gain control should be marked at the proper setting for each band, because the signal amplitude that gives the correct deviation on one band will be either too great or too small on another. For narrow-band FM the proper deviation is approximately 2000 cycles (based on an upper a.f. limit of 3000 cycles and a deviation ratio of 0.7) at the final *output* frequency. If the output frequency is in the 29-Mc. band and the oscillator is on 7 Mc., the deviation at the *oscillator* frequency should not exceed 2000/4, or 500 cycles.

Checking with a Crystal-Filter Receiver

With PM the d.c. method of checking just described cannot be used, because the frequency deviation at zero frequency also is zero. For narrow-band PM it is necessary to check the actual width of the channel occupied by the transmission. (The same method also can be used to check FM.) For this purpose it is necessary to have a crystal-filter receiver and an a.f. oscillator that generates a 3000-cycle sine wave.

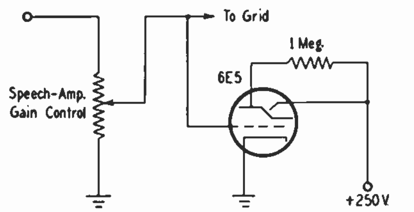


Fig. 9-58 — **6E5 modulation indicator for FM or PM modulators. To insure sufficient grid voltage for a good deflection, it may be necessary to connect the gain control in the modulator grid circuit rather than in an earlier speech-amplifier stage.**

Keeping the signal intensity in the receiver at a medium level, tune in the carrier at the *output* frequency. Do not use the a.v.c. Switch on the beat oscillator, and set the crystal filter at its sharpest position. Peak the signal on the crystal and adjust the b.f.o. for any convenient beat note. Then apply the 3000-cycle tone to the speech amplifier (use the connections shown in Fig. 9-41 to avoid overloading) and increase the audio gain until there is a small amount of modulation. Tuning the receiver on either side of the carrier frequency will show the presence of sidebands 3 kc. from the carrier on both sides. With low audio input, these two should be the only sidebands detectable.

Now increase the audio gain and tune the receiver over a range of about 10 kc. on both sides of the carrier. When the gain becomes high enough, a second set of sidebands spaced 6 kc. on either side of the carrier will be detected. The signal amplitude at which these sidebands become detectable is the maximum speech amplitude that should be used. If the 6E5 modulation indicator is incorporated in the modulator, its deflection with the 3000-cycle tone will be the "100-per-cent modulation" deflection for speech.

When this method of checking is used with a reactance-tube modulated FM (not PM) transmitter, the linearity of the system can be checked by observing the *carrier* as the a.f. gain is slowly increased. The beat-note frequency will stay constant so long as the modulator is linear, but nonlinearity will be accompanied by a shift in the average carrier frequency that will cause the beat note to change in frequency. If such a shift occurs at the same time that the 6-kc. sidebands appear, the extra sidebands may be caused by modulator distortion rather than by an excessive modulation index. This means that the modulator is not able to shift the frequency over a wide-enough range. The 6-kc. sidebands should appear *before* there is any shift in the carrier frequency.

R.F. Amplifiers

The r.f. stages in the transmitter that follow the modulated stage may be designed and adjusted as in ordinary operation. In fact, there are no special requirements to be met except that all tank circuits should be carefully tuned to resonance (to prevent unwanted r.f. phase shifts that might interact with the modulation and thereby introduce hum, noise and distortion). In neutralized stages, the neutralization should be as exact as possible, also to minimize unwanted phase shifts. With FM and PM, all r.f. stages in the transmitter can be operated at the manufacturer's maximum c.w.-telegraphy ratings, since the average power input does not vary with modulation as it does in AM 'phone operation.

The output of the transmitter should be checked for amplitude modulation by observing the antenna current. It should not change from the unmodulated-carrier value when the transmitter is modulated. If there is no antenna ammeter in the transmitter, a flashlight lamp and loop can be coupled to the final tank coil to serve as a current indicator. If the carrier amplitude is constant, the lamp brilliance will not change with modulation.

Amplitude modulation accompanying FM or PM is just as much to be avoided as frequency or phase modulation that accompanies AM. A mixture of AM with either of the other two systems results in the generation of spurious sidebands and consequent widening of the channel. If the presence of AM is indicated by variation of antenna current with modulation, the cause is almost certain to be nonlinearity in the modulator. In very wide-band FM the selectivity of the transmitter tank circuits may cause the amplitude to decrease at high deviations, but this is not likely to occur on amateur frequencies at which wide-band FM would be used.

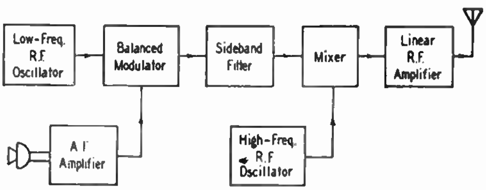
Single-Sideband Transmission

The most recent development in amateur radiotelephony is the introduction of practical single-sideband suppressed-carrier transmission. This system has tremendous potentialities for increasing the effectiveness of 'phone transmission and for reducing interference. Because only one of the two sidebands normally produced in modulation is transmitted, the channel width is immediately cut in half. However, when only one sideband is transmitted the carrier — which is essential in double-sideband transmission — no longer is necessary; it can be supplied without too much difficulty at the receiver. With the carrier eliminated there is a great saving in power at the transmitter — or, from another viewpoint, a great increase in effective power output. Assuming that the same final-amplifier tube or tubes are used either for normal AM or for single-sideband, carrier suppressed, it can be shown that the use of SSB gives an effective gain of at least 9 db. over AM — equivalent to increasing the transmitter power 8 times. Eliminating the carrier also eliminates the heterodyne interference that wrecks so much communication in congested 'phone bands.

Two basic systems for generating SSB signals are shown in Fig. 9-59. One involves the use of a bandpass filter having sufficient selectivity to pass one sideband and reject the other. Filters having such characteristics can only be constructed for relatively low frequencies, and most filters used by amateurs are designed to work somewhere between 10 and 20 kc. Good sideband filtering can be done at frequencies as high as 100 kc. by using multiple-crystal filters. The low-frequency oscillator output is combined with the audio output of a speech amplifier in a "balanced modulator" — one in which the carrier is "neutralized" out, and only the upper and lower sidebands appear in the output. One of the sidebands is passed by the filter and the other rejected, so that an SSB signal is fed to the mixer. The signal is there mixed with the output of a high-frequency r.f. oscillator to produce the desired output frequency. For additional amplification a linear r.f. amplifier (Class A or Class B) must be used. When the SSB signal is generated at 10 or 20 kc., it is generally first heterodyned to somewhere around 500 kc. and then to the operating frequency. This simplifies the problem of rejecting the "image" frequencies resulting from the heterodyne process.

The second system is based on the phase relationships between the carrier and sidebands in a modulated signal. As shown in the diagram, the audio signal is split into two components that are identical except for a phase difference of 90 degrees. The output of the r.f. oscillator (which may be at the operating frequency, if desired) is likewise split into two separate components having a 90-degree phase difference. One r.f. and one audio component are combined in each of two separate balanced modulators. The carrier is suppressed in the modulators, and the relative phases of the sidebands are such that one sideband is balanced out and the other is accentuated in the combined output. If the output from the balanced modulators is high enough, such an SSB exciter can work directly into the antenna, or the power level can be increased in a linear amplifier following the exciter.

Which is the better method of generating an SSB signal, the filter or the phasing method, is a controversial question. Properly adjusted, either system is capable of good results. Arguments in favor of the filter system are that it is somewhat easier to adjust without an oscilloscope, since it requires only a receiver and a v.t.v.m. for alignment, and it is more likely to remain in adjustment over a long period of time. The chief argument against it, from the amateur viewpoint, is that it requires quite a few stages and at least two frequency conversions after modulation. The phasing system requires fewer stages and can be designed to require no frequency conversions, but its alignment and adjustment are often considered to be a little "trickier" than that of the filter system. This probably stems from lack of familiarity with the system rather than any actual difficulty. In most cases the phasing system will cost less to apply to an existing transmitter.



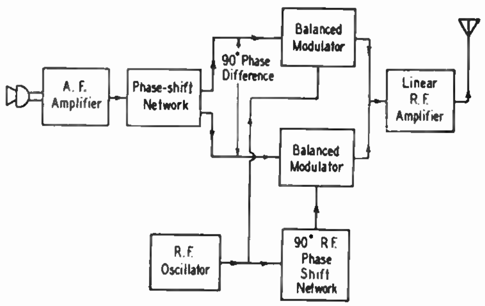


Fig. 9-59 — Two basic systems for generating single-sideband suppressed-carrier signals.

A One-Band SSB Exciter

The exciter shown in Figs. 9-60, 9-62 and 9-63 is an excellent unit for the amateur who might like to try single sideband with a minimum of cost and effort. It requires r.f. driving power from one's present exciter, audio power from an existing speech amplifier, and a power supply. It will deliver SSB output in the 3.9-Mc. 'phone band, either to an antenna for local work or to an r.f. amplifier adjusted for linear operation. The operating frequency can be varied over a wide range without seriously impairing the adjustment. Provision is made for transmitting either the upper or the lower sideband.

The complete schematic of the exciter is shown in Fig. 9-61. Four 6V6 tubes are used as balanced modulators. The plate circuit of the balanced modulators uses a push-pull-parallel arrangement. The grids of one pair of balanced modulators are fed through a phase-shift network consisting of a 300-ohm resistor and an inductance that is adjustable to 300 ohms reactance at the operating frequency. The grids of the second pair of balanced modulators are fed through a phase-shift network consisting of a 300-ohm resistor and a condenser which is adjustable to 300 ohms reactance at the operating frequency. The input impedance of the two phase-shift networks in parallel is 300 ohms. A grid-leak resistor, suitably by-passed, provides bias for each pair of balanced modulators.

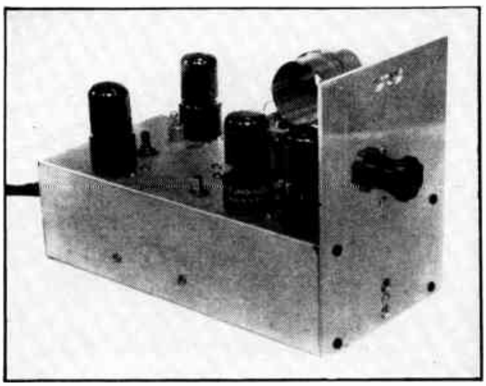


Fig. 9-60 — A small single-sideband exciter that can be used with practically any 75-meter 'phone rig. Receiving tubes are used. (W2UNJ, Aug., 1949, *QST*.)

The screen of each balanced-modulator tube is by-passed to ground for r.f. Screen modulation is used, and therefore each screen-dropping resistor is by-passed for audio. Two of the resistors are variable to allow balancing of the modulators.

A tapped audio inductance is used in the output of each audio amplifier, to provide push-pull modulating voltages from the single-ended amplifiers. A voltage divider is inserted between each output of the audio phase-shift network and the corresponding amplifier grid. One of these voltage dividers is made variable to provide for balancing of the two audio channels. The network constants are compensated for the load of these voltage dividers.

The Audio Phase-Shift Network

The audio phase-shift network requires close matching of resistance and capacity values and, to do this economically, advantage is taken of the fact that resistors and condensers in junk boxes and in stock at local dealers vary considerably from their nominal values.

Table 9-II is used in selecting the network components. The procedure is to collect as many resistors and condensers as possible with nominal values as indicated in the second column of the chart. Measure all of the condensers first, and select the six condensers whose measured values are closest to the "target values" in the third column. Enter the measured values of these condensers in the fourth column of the chart. Then calculate the "target values" for the resistors and select the six resistors whose measured values are closest to these target values.

A capacity bridge, of the type used by servicemen, and a good ohmmeter should give sufficient accuracy in selecting the network components. Absolute accuracy is not important, if the components are all in correct proportion to each other. A difference in percentage error between the resistance measurements and the capacitance measurements will merely shift the operating range of the network. The network components are mounted on a small sheet of insulating material to facilitate wiring.

TABLE 9-II
Phase-Shift Network Design Data

Part	*Nominal Value*	*Target Value*	*Measured Value*
C_1	0.001	0.00105	(C_{m1})
C_2	0.002	0.00210	(C_{m2})
C_3	0.006	0.00639	(C_{m3})
C_4	0.005	0.00475	(C_{m4})
C_5	0.01	0.00950	(C_{m5})
C_6	0.03	0.0285	(C_{m6})
R_1	100,000	$\frac{100}{C_{m1}} =$	
R_2	50,000	$\frac{105}{C_{m2}} =$	
R_3	15,000	$\frac{100}{C_{m3}} =$	
R_4	100,000	$\frac{453}{C_{m4}} =$	
R_5	50,000	$\frac{476}{C_{m5}} =$	
R_6	15,000	$\frac{453}{C_{m6}} =$	

All condensers mica, and all resistors 1 watt.

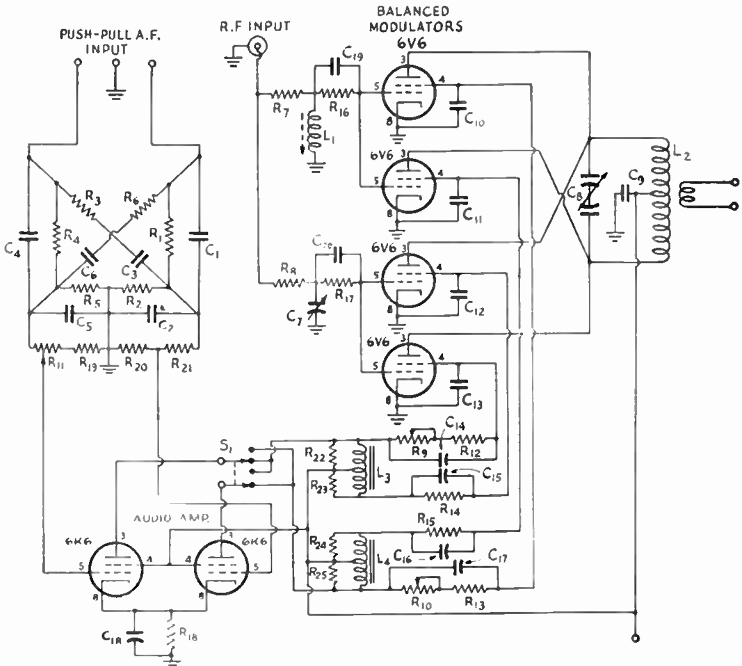


Fig. 9-61 — Circuit diagram of the single-sideband exciter.

C_1 – C_6 — See Table 9-II.
C_7 — 150-μμfd. air padder condenser.
C_8 — 100-μμfd.-per-section dual variable.
C_9 – C_{13}, C_{19}, C_{20} — 0.001-μfd. 500-volt mica.
C_{14} – C_{17} — 1-μfd. 150-v. electrolytic.
C_{18} — 10-μfd. 50-volt electrolytic.
R_1 – R_6 — See Table 9-II.
R_7, R_8 — Eight or more one-watt resistors of equal value paralleled to give 300 ohms.
R_9, R_{10} — 20,000-ohm potentiometers.
R_{11} — 0.5-megohm potentiometer.
R_{12}, R_{13} — 22,000 ohms, 1 watt.
R_{14}, R_{15} — 33,000 ohms, 1 watt.
R_{16}, R_{17} — 10,000 ohms, 1 watt.
R_{18} — 300 ohms, 2 watts.
R_{19} — 0.5 megohm, 1 watt.
R_{20} — 0.75 megohm, 1 watt.
R_{21} — 0.25 megohm, 1 watt.
R_{22}, R_{23}, R_{24}, R_{25} — 33,000 ohms, ½ watt.
L_1 — See text.
L_2 — Low-power 80-meter coil. (Bud OCL-80 with base removed. Six-turn link substituted for original.)
L_3, L_4 — Midget push-pull output to voice coil transformer (voice-coil winding not used).
S_1 — D.p.d.t. switch.

The R.F. Phasing Inductance

The only other "tricky" component of the exciter is the r.f. phasing inductance, L_1. This inductance is wound on a slug-tuned form salvaged from an i.f. transformer. The form is about three-eighths of an inch in diameter and one and five-eighths inches long. The winding is forty turns of No. 30 d.c.c. wire, close-wound. Since duplication of this inductance may be difficult, it is recommended that the constructor use a slug-tuned form and wire from his own junk box, and wind a coil that will resonate at 3.9 Mc. at the center of the slug-tuning range, with a variable condenser set at about 155 μμfd. Resonance can be checked by using the coil and condenser as a wavetrap connected in series with the antenna on the station receiver.

Construction

The exciter is assembled on a 5 × 10 × 3-inch chassis. The plate-tank tuning condenser is mounted on top of the chassis, front and center, with two of the 6V6 modulator tubes on each side. The plate tank coil is mounted on top of the condenser. Plate leads from the four 6V6s are brought directly to the tuning condenser through four ⅜-inch holes drilled through the chassis near each tube-socket plate connection. The 6V6 screen grids are by-passed to ground directly at the sockets. R_9, L_1, C_7 and R_{10} (all adjustable components) are mounted in a row directly behind the 6V6s. The two 6K6 amplifiers are mounted at the rear of the chassis, one on each side, with R_{11} and S_1 between them. The audio phase-shift network is mounted inside the chassis at the rear. Crystal sockets are used for r.f. input and output connections. A cable is brought out at the rear of the chassis for audio and power connections. Layout, construction and wiring are all conventional. The 5 × 7-inch front panel is optional.

Associated Equipment

The r.f. input impedance of the exciter is 300 ohms, but a link line of lower characteristic impedance will operate satisfactorily for the short distance usually required. A means for adjusting the r.f. driving power is desirable. A surplus Command set transmitter (BC-696 or

T-19/ARC-5), operating at low plate voltages, makes an ideal r.f. source, but any VFO or crystal oscillator with a few watts output will do.

In most stations, the handiest source of push-pull audio for the exciter will be the secondary of the modulator driver transformer. A single triode-connected 6F6 output tube in the speech amplifier will provide sufficient audio. The modulator tubes should be removed from their sockets, and the center tap of the driver-transformer secondary should be grounded, after removing the bias connection. An alternative method is to use blocking condensers in the audio leads to the single-sideband exciter to isolate the modulator bias from the audio phase-shift network in the exciter. If some other source of push-pull audio is used, it should have low internal impedance (Class A triodes, or beam tubes with negative *voltage* feed-back).

Operating Conditions

The operating conditions for the exciter are determined by the required output. If the required output is low, it is better to run the exciter with low plate voltages. This will reduce the amount of residual carrier present in the output in relation to the sideband output. Also, the exciter will be more stable and maintain adjustment longer with lower plate voltages.

The power input to the modulator plates should not exceed 30 watts with no audio input. The input to the modulators may be varied by adjusting the voltage used on the amplifiers and modulator screens.

The exciter may be coupled directly to an antenna for use as a low-power transmitter, but most amateurs will wish to use it to drive a buffer or final amplifier. All stages following the exciter must be operated under Class A, AB, or B conditions. In general, the correct operating conditions for stages following the exciter may be found by referring to the audio operating conditions for the tube under consideration. Grid-bias and screen voltages should have very good regulation. For amateur voice operation, tubes may be operated considerably beyond the ratings given in the tube manuals, as discussed later. When the r.f. amplifier is operated Class AB_2 or Class B, the grid tank circuit should be shunted by a resistor in order to provide better regulation of the exciting voltage. The value of this resistor is not critical and may be determined by experiment.

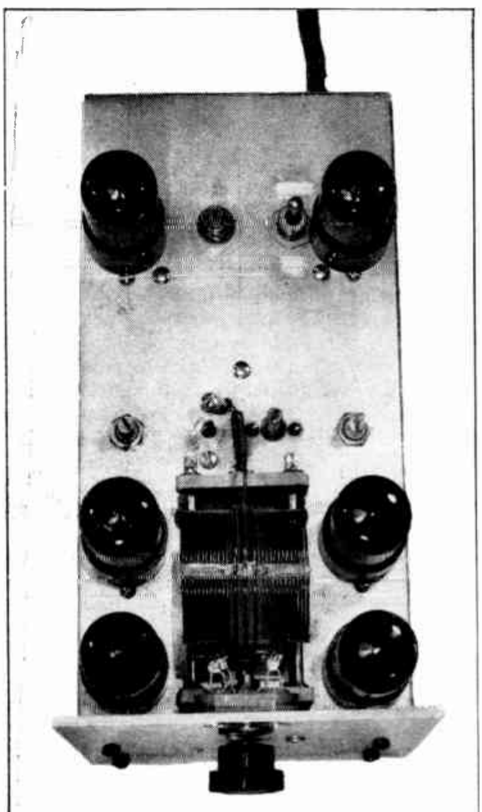


Fig. 9-62 — A top view of the exciter. The toggle switch at the rear selects the sideband in use.

Adjustment

Adjustment of the exciter is best made under actual operating conditions. Connect the exciter to the transmitter, load the transmitter into a dummy load, apply r.f. excitation to the exciter, feed a source of sine-wave audio into the speech amplifier, and tune the transmitter in the conventional way for maximum output.

Reduce the audio input to zero, and adjust potentiometers R_9 and R_{10} for minimum carrier output. Minimum carrier output may be determined by any sensitive r.f. indicator coupled to the final-amplifier plate circuit. A 0–1 milliammeter, in series with a crystal detector and a two-turn coupling loop, will make a satisfactory indicator. The meter should be by-passed with a 0.005-μfd. condenser. If a null indication cannot be obtained within the range of the potentiometers, the 6V6 tubes are not evenly matched. Exchanging the positions of the 6V6s may aid in obtaining the balance, or other tubes may have to be used.

After the carrier balance is obtained, tune in the r.f. source on the station receiver, and with the antenna terminals shorted, and the crystal selectivity in sharp position, adjust the crystal phasing to the point where only one sharply-peaked response is obtained as the receiver is tuned through the signal. Now apply sine-wave audio of about 1500-cycle frequency to the speech amplifier, and find the two sidebands on the receiver. Three distinct peak indications will be observed on the S-meter as the receiver is tuned. Set the receiver on the weaker of the two sidebands and adjust L_1, C_7 and R_{11} for

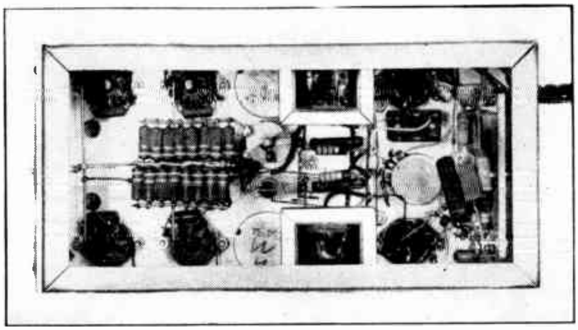


Fig. 9-63 — A bottom view of the exciter. The phase-shift network is mounted on a panel at the right-hand side. The double string of resistors at the left is the load for the r.f. excitation.

minimum sideband strength. If suppression of the other sideband is desired, throw S_1 to its other position. A dip obtained with one set of adjustments is not necessarily the minimum. Other combinations should be tried. The final adjustment should give S-meter readings for the two sidebands which differ by at least 30 db. The bias voltages on the two pairs of balanced modulators will be equal.

After the adjustments have been completed, the r.f. drive to the exciter should be adjusted to the point where a decrease in drive will cause a decrease in output, but an increase in drive will not cause an increase in output. The complete adjustment procedure should then be rechecked. The rig is then ready for a microphone, an antenna, and an on-the-air test.

If an oscilloscope is available, a simpler and more reliable adjustment procedure may be used. Either linear or sine-wave horizontal sweep may be used on the oscilloscope. The vertical input should be coupled to the output of the transmitter in the same manner as is used for observing amplitude modulation. The sine-wave audio-frequency input to the speech amplifier should be any convenient multiple of the oscilloscope sweep frequency. A 60-cycle sweep frequency and a 600-cycle audio frequency are commonly used.

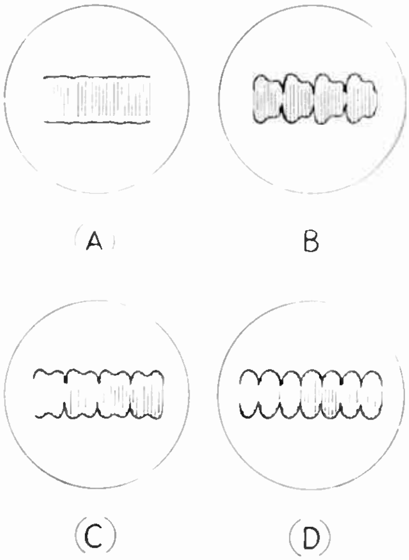


Fig. 9-64 — Sketches of the oscilloscope face showing different conditions of adjustment of the exciter unit. (A) shows the substantially clean carrier obtained when all adjustments are at optimum and a sine-wave signal is fed to the audio input. (B) shows improper r.f. phase and unbalance between the outputs of the two balanced modulators. (C) shows improper r.f. phasing but outputs of the two balanced modulators equal. (D) shows proper r.f. phasing but unbalance between outputs of two balanced modulators.

When the exciter is modulated with a single sine-wave audio frequency, the output should be a single radio frequency. Therefore, the oscilloscope should show a straight-edged band across the screen, the same indication as is given by an unmodulated carrier. This is illustrated in Fig. 9-64A. If carrier output, or unwanted sideband output, is present, it will be indicated by "ripple" on the top and bottom edges of the oscilloscope picture. A small amount of ripple can be tolerated, but if the exciter is badly out of adjustment, the output will appear to be heavily modulated. Adjustment with the 'scope is accomplished by adjusting all controls to obtain the smallest possible amount of ripple. The oscilloscope may also be used for continuous monitoring during transmissions to avoid overloading of any stage of the transmitter. Overloading is indicated by a flattening of the modulation-peak patterns at the top and bottom. In observing these patterns, it is difficult to separate the effects of sideband and carrier suppression. However, considered separately, sideband or carrier supression of 30 db. would give a 3 per cent ripple, 25 db. a ripple of 6 per cent, and 20 db. a 10 per cent ripple. Harmonics present in the audio modulating signal will modify the results and invalidate this test if they run more than 1 per cent.

A pair of 807s operating Class AB_2 can be driven by the exciter with only 60 ma. (at 120 volts) input to the balanced modulators, and with the exciter amplifiers also operating at 120 volts. Part of the output of the exciter is, of course, dissipated in the load resistor across the grid tank circuit of the 807s. The balanced modulators require sufficient r.f. drive to develop 12 volts of grid bias under these operating conditions.

Amplification of SSB Signals

When an SSB signal is generated at some frequency other than the operating frequency, it is necessary to change frequency by heterodyne methods. These are exactly the same as those used in receivers, and any of the normal mixer or converter circuits can be used. One exception to this is the case where the original signal and the heterodyning oscillator are not too different in frequency (as when heterodyning a 20-kc. signal to 500 kc.) and, in this case, a balanced mixer should be used, to eliminate the heterodyning oscillator frequency in the output and thus reduce the chances for spurious signals appearing in the output.

To increase the power level of an SSB signal, a "linear amplifier" must be used. The simplest form of linear amplifier (r.f. or audio) is the Class A amplifier, which is used almost without exception throughout our receivers and our low-level speech equipment. While its linearity can be made phenomenally good, it is unfortunately quite inefficient. The theoretical limit of efficiency in this case is 50 per cent, while most practical amplifiers run 25–35 per cent efficient at full output. At low levels this is not worth worrying about, but when the 2- to 10-watt level is exceeded something else must be done to improve this efficiency and reduce tube, power-supply and operating costs.

Class B amplifiers are theoretically capable of 78.5 per cent efficiency at full output, and practical amplifiers run at 60–70 per cent efficiency at full output. Tubes normally designed for Class B audio work can be used in r.f. linear amplifiers and will operate at the same power rating and efficiency provided, of course, that the tube is capable of operation at the radio frequency. The operating conditions for r.f. are substantially the same as for audio work — the only difference is that the input and output transformers are replaced by suitable r.f. tank circuits. Further, in r.f. circuits it is readily possible to operate only one tube if only half the power is wanted — push-pull is not a necessity in Class B r.f. work. However, the r.f. harmonics will be higher in the case of the single-ended amplifier, and this should be taken into consideration if TVI is a problem.

In a few instances, Class B r.f. amplifier ratings of tubes are given in the tube books, and the efficiency shown will be about 33 per cent. These ratings are for use when carrier is present and do not apply to SSB suppressed-carrier operation. The Class B audio ratings are a better indication of what can be expected.

For proper operation of Class B amplifiers, and to reduce harmonics and facilitate coupling, the input and output circuits should not have a low C-to-L ratio. A good guide to the proper size of tuning condenser is the chart of Fig. 6-22 and, in case of any doubt, it is well to be on the high-capacity side. If zero-bias tubes are used in the Class B stage, it will not be necessary to add much "swamping" resistance across the grid circuit, because the grids of the tubes load the circuit at all times. However, with other tubes that require bias, the swamping resistor should be such that it dissipates from five to ten times the power required by the grids of the tubes. This will insure an almost constant load on the driver stage and good regulation of the grid voltage of the Class B stage.

Before going into detail on the adjustment and loading of the Class B linear amplifier, a few general considerations should be kept in mind. If proper operation is expected, it is essential that the amplifier be so constructed, wired and neutralized that no trace of regeneration or parasitic instability remains. Needless to say, this also applies to the stages driving it.

The bias supply to the Class B linear amplifier should be quite stiff. A Class C stage thrives on grid-leak bias, but for really good operation the Class B should be supplied from a very stiff source, such as batteries or some form of voltage regulator. If nonlinearity is noticed when testing the unit, the bias supply may be checked by means of a large electrolytic capacitor. Simply shunt the supply with 100 μfd. or so of capacity and see if the linearity improves. If so, rebuild the bias supply for better regulation. *Do not rely on a large condenser alone.*

Adjustment of Amplifiers

The two critical adjustments for obtaining proper operation from the linear amplifier that has been correctly designed are the plate loading and the grid drive. Since these adjustments are preferably made with power on, it is a matter of practical convenience to have both controls readily available, at least during initial tune-up.

The 'scope can show misadjustment at a glance and will greatly facilitate all adjustments. In addition, it is the most reliable instrument for observing modulation amplitude

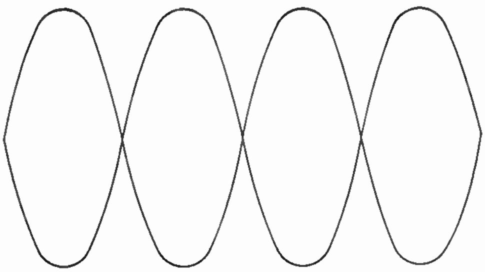


Fig. 9-65 — Oscilloscope pattern obtained with a two-tone test signal through a correctly-adjusted linear amplifier.

and, once used, is likely to become the most nearly essential instrument in the shack. Nothing elaborate is needed.

With single sideband, 100 per cent modulation with a single tone is a pure r.f. output with no modulation envelope, and the point of amplifier overload is difficult to observe. However, if the input signal consists of two sine waves of different frequencies (for example, 1000 c.p.s. difference) but equal amplitudes, the output of the single-sideband transmitter should have the envelope shown in Fig. 9-65. This is called a "two-tone" test signal to distinguish it from other test signals. Its first advantage lies in the fact that any flattening of the positive peaks is readily discernible, which makes the adjustment of the linear-amplifier drive and output coupling as simple a procedure as that for AM systems.

Those who use the filter method for obtaining single-sideband signals can obtain such a test signal by mixing the output of two audio oscillators of good waveform. The experimenters using the phasing method of single-sideband signal generation will recognize the pattern as that obtained when a single test tone is applied to one of their balanced modulators. For this latter group a two-tone test signal may be readily obtained by disabling one of the balanced modulators in the exciter and applying a single input tone. Other variations are possible in different exciters, and the final choice of any one operator will be dictated by convenience.

Suppose that the linear amplifier has been coupled to a dummy load and the single-sideband exciter has been connected to its input. By observing the oscilloscope coupled to the amplifier output, it will be possible to adjust the drive and output coupling so that the peaks of the two-tone test signal waveform are on the verge of flattening. The peak input power may now be checked. This is readily possible, for with the two-tone test signal applied, the peak input power will be 1.57 times the d.c. power input to the linear amplifier. Should this be different from the design value for the particular linear amplifier, the drive and loading adjustments can be quickly changed in the proper direction (always adjusting the loading so that the peaks of the envelope are on the verge of flattening) and the proper design value reached.

As a final check, before coupling the linear amplifier to the antenna, the single-sideband operator will do well to check the linearity of the system, since distortion in the linear amplifier (for that matter, in any of the r.f. amplifiers) probably will result in the generation of sidebands on the side that was suppressed in the exciter. Here again the two-tone test signal will be of great help, since distortion of the signal will be readily recognized. A check of the bias supply has already been recommended. The next most likely form of distortion will be caused by curvature of the tube characteristic near cut-off, and will be recognizable from a two-tone test pattern that looks like Fig. 9-66. A slight readjustment of bias (or applying a few volts of positive or negative bias, in the case of zero-bias tubes) will usually straighten out the kink that exists where the pattern crosses the zero axis. Make this adjustment with special care, however, because the dissipation of the tubes with no input signal will be very sensitive to this adjustment. There are a few tubes that will not permit this adjustment to be carried to the point where the kink is entirely eliminated without exceeding the rated plate dissipation.

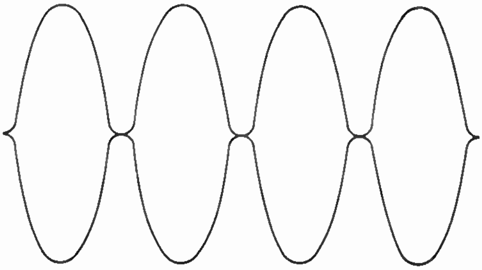


Fig. 9-66 — **The distorted two-tone test-signal pattern obtained when the bias voltage is incorrect.**

The antenna may now be coupled to the linear amplifier until the plate input with the excitation as determined above is the same as that obtained with the dummy load. The operator can now feel that the system has been adjusted for optimum performance.

Further details and recent developments in amateur equipment for single-sideband work will be found in the following references:

Goodman, "What Is Single-Sideband Telephony?" *QST*, January, 1948.

Villard, "Single-Sideband Operating Tests," *QST*, January, 1948.

Nichols, "A Single-Sideband Transmitter for Amateur Operation," *QST*, January, 1948.

Norgaard, "What About Single Sideband?" *QST*, May, 1948.

Norgaard, "A New Approach to Single Sideband," *QST*, June, 1948.

Dawley, "An S.S.S.C. Transmitter Adapter," *QST*, July, 1948.

Villard, "A Simple Single-Sideband Transmitter," *QST*, November, 1948.

Goodman, "The Basic 'Phone Exciter," *QST*, January, 1949; correction ("Feed-back"), March, 1949.

Mann, "An Inexpensive Sideband Filter," *QST*, March, 1949.

Reque, "Linear R.F. Amplifiers," *QST*, May, 1949.

Berry, "Filter Design for the Single-Sideband Transmitter," *QST*, June, 1949.

Rust, "Single Sideband for the Average Ham," *QST*, August, 1949.

Goodman, "A 75- and 20-Meter Single-Sideband Exciter," *QST*, November, 1949.

Nibbe, "Audio Phase-Shift Networks," *QST*, January, 1950.

CHAPTER 10

Antennas and Transmission Lines

The radio-frequency power that is generated by a transmitter serves a useful purpose only when it is radiated out into space in the form of electromagnetic waves. It is the antenna's job to convert the power into radio waves as efficiently as possible, and to direct the waves where they will do the most good in communication. To do so, the antenna usually must be located well above the ground and kept as far as possible from buildings, trees, and other objects that might absorb energy. This raises a problem, because by some means or another the power that is generated in the transmitter must be conveyed to the antenna. The usual means is a **transmission line.**

There is thus a natural association between antennas and transmission lines — an association that has frequently led to the quite mistaken belief that an antenna fed by a particular type of transmission line is a better (or worse) radiator than exactly the same type of antenna fed by a different type of transmission line. The fact is that a transmission line can be used to carry power to any sort of device — not just an antenna — capable of receiving it. Nor does the antenna care by what means it gets the power; the amount it receives will be radiated just as well no matter by what system it was conveyed to the antenna proper.

Transmission Lines

Suppose we have a battery and a pair of parallel wires extending to a very great distance. At the moment the battery is connected to the wires, electrons in the wire near the positive terminal will be attracted to the battery, and the same number of electrons in the wire near the negative battery terminal will be repelled outward along the wire.

Thus a current flows in both wires near the battery at the instant the battery is connected. However, a definite time interval will elapse before these currents are evident at a distance from the battery. The time interval may be very small. For example, one-millionth of a second (one microsecond) after the connection is made the currents in the wires will have traveled 300 meters, or nearly 1000 feet, from the battery terminals.

The current is in the nature of a charging current, flowing to charge the capacitance between the two wires. But unlike an ordinary condenser, the conductors of this "linear" condenser have appreciable inductance. In fact, we may think of the line as being composed of a whole series of small inductances and capacitances connected as shown in Fig. 10-1, where each coil is the inductance of a very short section of one wire and each condenser is the capacitance between two such short sections.

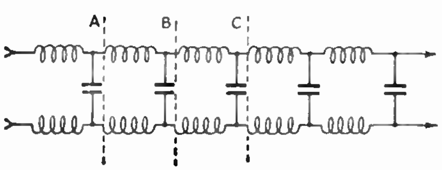


Fig. 10-1 — Equivalent of a transmission line in lumped circuit constants.

Characteristic Impedance

An infinitely-long chain of coils and condensers connected as in Fig. 10-1, where each L is the same as all others and all the Cs have the same value, has an important peculiarity. To an electrical impulse applied at one end, the combination appears to have an impedance — called the **characteristic impedance** or **surge impedance** — that is approximately equal to $\sqrt{L/C}$. This impedance is purely resistive. In a transmission line, L and C are the inductance and capacitance per unit length.

The inductance and capacitance per unit length depend upon the size of the line conductors and the spacing between them. The closer the two conductors of the line and the greater their diameter, the higher the capacitance and the lower the inductance. A line with large conductors closely spaced will have low impedance, while one with small conductors widely spaced will have relatively high impedance.

The characteristic impedance of the line determines the amount of current that can flow when a voltage is applied to the line. When a line is infinitely long, the current is simply equal to E/Z_0, where E is the voltage applied to the line and Z_0 is the characteristic impedance. This has nothing to do with the *resistance* of the conductors. The line is an impedance (like any circuit composed of L and C, without any R) that does not consume power. Actually, of course, the conductors do have resistance, so power cannot be transmitted along the line without some loss. But if the line is properly constructed and operated, this loss will be small compared with the amount of power carried to the load to do useful work.

"Matched" Lines

In this picture of current traveling along a transmission line we have assumed that the line was infinitely long. Actual lines have a definite length and are connected to or **terminated** in a load at the "output" end, or end to which the power is delivered. If the load is a pure resistance of a value equal to the characteristic impedance of the line, the current traveling along the line to the load does not find conditions changed in the least when it meets the load; in fact, the load just looks like still more transmission line of the same characteristic impedance. Consequently, connecting such a load to a short transmission line allows the current to travel in exactly the same fashion as it would on an infinitely-long line.

In other words, a short line terminated in a purely-resistive load equal to the characteristic impedance of the line acts just as though it were infinitely long. Such a line is said to be **matched.** In a matched transmission line, power travels outward along the line from the source until it reaches the load, where it is completely absorbed.

R.F. on Lines

The discussion above, although based on direct-current flow from a battery, also holds when an r.f. voltage is applied to the line. The difference is that the alternating voltage causes the amplitude of the current at the input terminals of the line to vary in the same way as the amplitude of the voltage, and the direction of current flow also periodically reverses when the polarity of the applied voltage reverses. In the time of one cycle the current will travel a distance of one wavelength along the line wires. Because the current at a given instant at any point along the line is the result of a voltage that was applied at some *earlier* instant at the input terminals, the instantaneous amplitude of the current is different at all points in a one-wavelength section of line; in fact, the current flows in opposite directions in the same wire in adjacent half-wavelength sections. This is the *instantaneous* picture. In contrast, at any given point along the line the current goes through the same variations in the time of one cycle that the current at the input terminals did.

The result of all this is that the current (and voltage) travels along the wire as a series of waves having a length equal to the velocity at which the current travels divided by the frequency of the a.c. voltage. On an infinitely-long line, or one properly matched at the load, an ammeter inserted anywhere in the line will show the same current, since the ammeter averages out the variations in current during a cycle. It is only when the line is not properly matched that the wave motion becomes apparent. This is discussed in the next section.

STANDING WAVES

With the infinitely-long line (or its matched counterpart) the impedance was the same at any point on the line and therefore the ratio of voltage to current was the same at any point on the line. However, the impedance at the end of the line in Fig. 10-2 is zero — or at least extremely small — because the line is short-circuited at the end. A given amount of power in a very low impedance will result in a very large current and a very small voltage, as compared with the current-voltage ratio that exists in a few hundred ohms — which is a typical impedance value for some types of transmission lines. Something has to happen, therefore, when the power traveling along the transmission line meets the short-circuit at the end.

What happens is that the outgoing power, on meeting the short-circuit, simply reverses its direction of flow and goes back along the transmission line toward the input end. It has nowhere else to go. There is a very large current in the short-circuit, but substantially no voltage across the line at this point. We now have a voltage and current representing the power going outward toward the short-circuit, and a second voltage and current representing the **reflected** power traveling back toward the source.

Consider only the two current components for the moment. The reflected current travels at the same speed as the outgoing current, so its instantaneous value will be different at every point along the line, in the distance represented by the time of one cycle. At some points along the line the phase of the outgoing and reflected currents will be such that the currents cancel each other while at others the amplitude is doubled. At in-between points the amplitude is between these two extremes. The points at which the currents are in and out of phase depend only on the *time* required for them to travel and so depend only on the *distance* along the line from the point of reflection. The phase is completely reversed in the time of one-half cycle —

that is, a distance of one-half wavelength — and is back in the in-phase condition when the current has traveled for one whole cycle, or one wavelength.

In the short-circuit at the end of the line the total current is high and the two current components are in phase. Therefore at a distance of *one-half* wavelength back along the line from the short-circuit the outgoing and reflected components will again be in phase and the current will have its maximum value. This is also true at any point that is a multiple of a half-wavelength from the short-circuited end of the line. The distance along the line is one-half wavelength, rather than a full wavelength, because the two components are traveling in opposite directions.

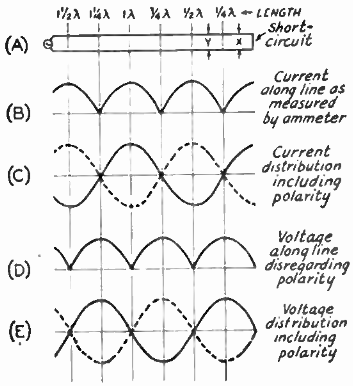


Fig. 10-2 — **Standing waves of voltage and current along a short-circuited transmission line.**

Since a total distance of one-half wavelength gives a complete reversal of phase, the outgoing and reflected currents will cancel at a point *one-quarter* wavelength, along the line, from the short-circuit. At this point, then, the current will be zero. It will also be zero at all points that are an *odd* multiple of one-quarter wavelength from the short-circuit.

If the current along the line is measured at successive points with an ammeter, it will be found to vary about as shown in Fig. 10-2B. The same result would be obtained by measuring the current in either wire, since the ammeter cannot measure phase. However, if the phase could be checked, it would be found that in each successive half-wavelength section of the line the currents at any given instant are flowing in opposite directions, as indicated by the solid line in Fig. 10-2C. Furthermore, the current in the second wire is flowing in the opposite direction to the current in the adjacent section of the first wire, as a result of the electron movement discussed earlier. This is indicated by the broken curve in Fig. 10-2C. The variations in current intensity along the transmission line are referred to as **standing waves.** The point of maximum line current is called a **current loop** and the point of minimum line current a **current node.**

Voltage Relationships

Since the end of the line is short-circuited, the voltage at that point has to be zero. This can only be so if the voltage in the outgoing wave is met, at the end of the line, by an equal voltage of opposite polarity. In other words, the phase of the voltage wave is *reversed* when reflection takes place from the short-circuit. This reversal is equivalent to an extra half-cycle or half-wavelength of travel. As a result, the outgoing and returning voltages are in phase a quarter wavelength from the end of the line, and again out of phase a half-wavelength from the end. The standing waves of voltage, shown at D in Fig. 10-2, are therefore displaced by one-quarter wavelength from the standing waves of current. The drawing at E shows the voltages on both wires when phase is taken into account. The polarity of the voltage on each wire reverses in each half-wavelength section of transmission line. A voltage maximum on the line is called a **voltage loop** and a voltage minimum is called a **voltage node.**

Input Impedance

It is apparent, from examination of B and D in Fig. 10-2, that at points that are a multiple of a half-wavelength — i.e., ½, 1, 1½ wavelengths, etc. — from the short-circuited end of the line the current and voltage have the same values that they do at the short-circuit. In other words, if the line were an exact multiple of a half-wavelength long the generator or source of power would "look into" a short-circuit. On the other hand, at points that are an odd multiple of a quarter wavelength — i.e., ¼, ¾, 1¼, etc. — from the short-circuit the voltage is maximum and the current is zero. Since $Z = E/I$, the impedance at these points is theoretically infinite. (Actually it is very high, but not infinite. This is because the current does not actually go to zero when there are losses in the line. Losses are always present, but usually are small enough so that the impedance is of the order of tens or hundreds of thousands of ohms.)

At either the odd or even multiples of a quarter wavelength the impedance is a pure resistance, because at these points the current and voltage in the transmission line are exactly in phase.

A detailed study of the outgoing and reflected components of voltage and current will show that at a point such as X in Fig. 10-2, lying anywhere in the section of line between the short-circuit and the first quarter-wavelength point, the current lags behind the voltage. This is exactly what happens in an inductance, so it can be said that a section of short-circuited transmission line less than a quarter wavelength long has inductive reactance. The value of reactance is determined

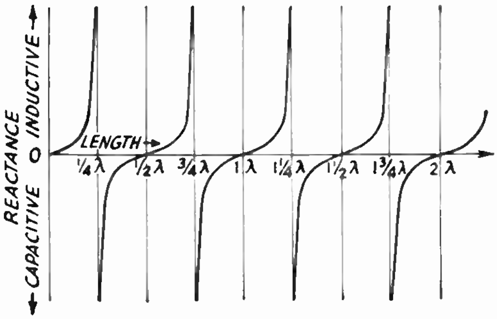


Fig. 10-3 — Input reactance *vs.* length of a short-circuited transmission line. Actual values of reactance depend upon the characteristic impedance of the line as well as its length. For a given line length, the input reactance is directly proportional to the characteristic impedance.

by the ratio of voltage to current at the input end of such a line. It is evident from B and D in Fig. 10-2 that the reactance is low when the line is quite short, and highest when the line is nearly a quarter wavelength long. The line also has inductive reactance when its length is between one-half and three-quarter wavelength, between one and one-and-one-quarter wavelengths, and so on.

On the other hand, in the section of line between one-quarter and one-half wavelength from the short-circuit, such as at Y in Fig. 10-2, the current leads the voltage. A short-circuited line having a length between these two limits "looks like" a capacitive reactance to the generator to which it is connected. The reactance is highest when the line is just over one-quarter wavelength long, and lowest when the line is just less than one-half wavelength long. Fig. 10-3 shows the general way in which the reactance varies with different line lengths.

Open-Circuited Line

If the end of the line is open-circuited instead of short-circuited, there can be no current at

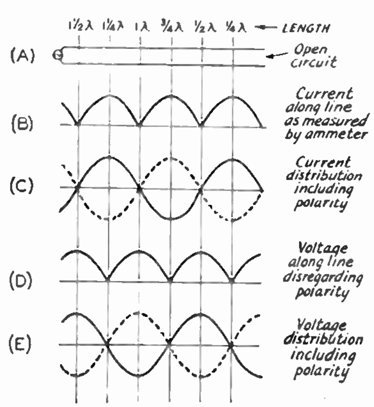


Fig. 10-4 — Standing waves of current and voltage along an open-circuited transmission line.

the end of the line but a large voltage can exist. Again the outgoing power is reflected back toward the source because it has nowhere else to go. In this case, the outgoing and reflected components of *current* must be equal and opposite in phase in order for the total current at the end of the line to be zero. The outgoing and reflected components of voltage are in phase, however, and add together. The result is that we again have standing waves, but the conditions are reversed. Fig. 10-4 shows the open-circuited line case. It may be compared directly with Fig. 10-2. The impedance looking into the line toward the open end is purely resistive at each multiple of one-quarter wavelength. It is very low at odd multiples of one-quarter wavelength, and very high at even multiples. In fact, an open-circuited line and short-circuited line behave just alike *if* the length of one differs by one-quarter wavelength from the length of the other.

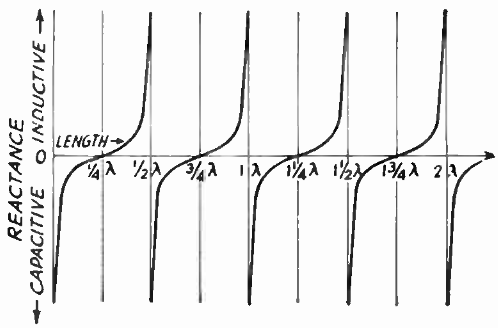


Fig. 10-5 — Input reactance *vs.* length of an open-circuited transmission line. Actual values of reactance depend upon the characteristic impedance of the line as well as its length. For a given line length, the input reactance is directly proportional to the characteristic impedance.

Fig. 10-5 shows how the reactance varies with line length for the open-circuited line. Comparing this with Fig. 10-3 shows that the reactance of any given length of line is of the opposite type to that obtained with a short-circuited line of the same length.

Lines Terminated in Resistive Load

Fig. 10-6 shows a line terminated in a resistive load. In such a case at least part of the outgoing power is absorbed in the load, and so is not available to be reflected back toward the source. Because only part of the power is reflected, the reflected components of voltage and current do not have the same magnitude as the outgoing components. Therefore there is no such thing as complete cancellation of either voltage or current at any point along the line. However, the *speed* at which the outgoing and reflected components travel is not affected by their amplitude, so the phase relationships are similar to those in open- or short-circuited lines.

It was pointed out earlier that if the load

Practical Line Characteristics

The foregoing discussion of transmission lines has been based on a line consisting of two parallel conductors. Actually, the **parallel-conductor** line is but one of two general types. The other is the **coaxial** or **concentric** line. The coaxial line consists of a round conductor placed in the center of a circular tube. The inside surface of the tube and the outside surface of the smaller inner conductor form the two conducting surfaces of the line.

In the coaxial line the fields are entirely inside the tube, because the tube acts as a shield to prevent them from appearing outside. This reduces radiation to the vanishing point. So far as the electrical behavior of coaxial lines is concerned, all that has previously been said about the operation of parallel-conductor lines applies. There are, however, practical differences in their construction and use.

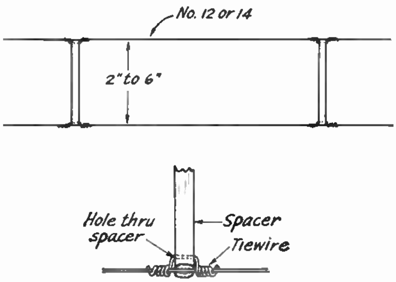


Fig. 10-8 — Typical construction of open-wire line. The line conductor fits in a groove in the end of the spacer, and is held in place by a tie-wire anchored in a hole near the groove.

Types of Construction

There are several constructional variations in both the basic types of transmission lines mentioned in the preceding section. Probably the most common type of transmission line used in amateur installations is a parallel-conductor line in which two wires (ordinarily No. 12 or No. 14) are supported a fixed distance apart by means of insulating rods called "spacers." The spacings used vary from two to six inches, the smaller spacings being necessary at frequencies of the order of 28 Mc. and higher so that radiation will be minimized. The construction is shown in Fig. 10-8. Such a line is said to be **air-insulated.** Typical spacers are shown in Fig. 10-9. The characteristic impedance of such "open-wire" lines runs between about 400 and 600 ohms, depending on the wire size and spacing.

Parallel-conductor lines also are sometimes constructed of metal tubing of a diameter of ¼ to ½ inch. This reduces the characteristic impedance of the line. Such lines are mostly used as quarter-wave transformers, when different values of impedance are to be matched.

Two forms of "Twin-Lead" or "ribbon" transmission line are shown in Fig. 10-9. This is a parallel-conductor line with stranded conductors imbedded in low-loss insulating material (polyethylene). It has the advantages of light weight, compactness and neat appearance, together with close and uniform spacing. However, losses are higher in the solid dielectric than in air, and dirt or moisture on the line tends to change the characteristic impedance. Twin-Lead line is available in characteristic impedances of 75, 150 and 300 ohms.

The most common form of coaxial line consists of either a solid or stranded-wire inner conductor surrounded by polyethylene dielectric. Copper braid is woven over the dielectric to form the outer conductor, and a waterproof vinyl covering is placed on top of the braid. This cable is made in a number of different diameters. It is moderately flexible, and so is convenient to install. Some different types are shown in Fig. 10-9. This solid coaxial cable is commonly available in impedances approximating 50 and 70 ohms.

Air-insulated coaxial lines have lower losses than the solid-dielectric type, but are less used in amateur work because they are expensive and difficult to install as compared with the flexible cable. The common type of air-insulated coaxial line uses a solid-wire conductor inside a copper tube, with the wire held in the center of the tube by means of insulating "beads" at regular intervals.

Characteristic Impedance

The characteristic impedance of an air-insulated parallel-conductor line is given by:

$$Z_0 = 276 \log \frac{b}{a} \qquad \textbf{(10-D)}$$

where Z_0 = Characteristic impedance
b = Center-to-center distance between conductors
a = Radius of conductor (in same units as b)

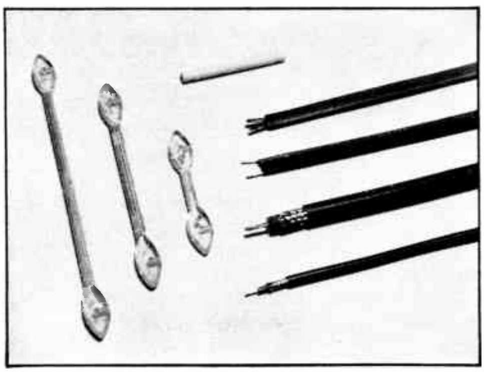


Fig. 10-9 — Typical manufactured transmission lines and spacers.

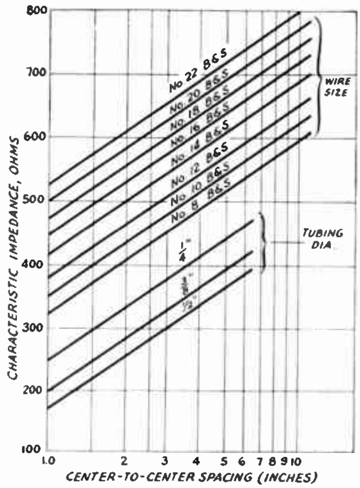


Fig. 10-10 — Chart showing the characteristic impedance of typical spaced-conductor parallel transmission lines. Tubing sizes given are for outside diameters.

It does not matter what units are used for a and b so long as they are the *same* units. Both quantities may be measured in centimeters, inches, etc. Since it is necessary to have a table of common logarithms to solve practical problems, the solution is given in graphical form in Fig. 10-10 for a number of common conductor sizes.

The characteristic impedance of an air-insulated coaxial line is given by the formula

$$Z_0 = 138 \log \frac{b}{a} \qquad \textbf{(10-E)}$$

where Z_0 = Characteristic impedance
b = Inside diameter of outer conductor
a = Outside diameter of inner conductor (in same units as b)

Again it does not matter what units are used for b and a, so long as they are the same. Curves for typical conductor sizes are given in Fig. 10-11.

The formula for coaxial lines is approximately correct for lines in which bead spacers are used, provided the beads are not too closely spaced. When the line is filled with a solid dielectric, the characteristic impedance as given by the chart should be multiplied by $1/\sqrt{K}$, where K is the dielectric constant of the material. In solid-dielectric parallel-conductor lines such as Twin-Lead the characteristic impedance cannot be calculated readily, because part of the electric field is in air as well as in the solid dielectric.

Electrical Length

In the discussion of line operation earlier in this chapter it was assumed that currents traveled along the conductors at the speed of light. Actually, the velocity is somewhat less, the reason being that electromagnetic fields travel more slowly in dielectric materials than they do in free space. In air the velocity is practically the same as in empty space, but a practical line always has to be supported in some fashion by solid insulating materials. The result is that the fields are slowed down; the currents travel a shorter distance in the time of one cycle than they do in space, and so the wavelength along the line is less than the wavelength would be in free space at the same frequency. (Wavelength is equal to velocity divided by frequency.)

Whenever reference is made to a line as being so many wavelengths (such as a "half-wavelength" or "quarter wavelength") long, it is to be understood that the *electrical* length of the line is meant. Its actual physical length as measured by a tape always will be somewhat less. The physical length corresponding to an electrical wavelength is given by

$$\textit{Length in feet} = \frac{984}{f} \cdot V \qquad \textbf{(10-F)}$$

where f = Frequency in megacycles
V = Velocity factor

The **velocity factor** is the ratio of the actual velocity along the line to the velocity in free space. Values of V for several common types of lines are given in Table 10-I.

> Example: A 75-foot length of 300-ohm Twin-Lead is used to carry power to an antenna at a frequency of 7150 kc. From Table 10-I, V is 0.82. At this frequency (7.15 Mc.) a wavelength is
>
> $$\textit{Length} \text{ (feet)} = \frac{984}{f} \cdot V = \frac{984}{7.15} \times 0.82 = 137.6 \times 0.82 = 112.8 \text{ ft.}$$
>
> The line length is therefore 75/112.8 = 0.665 wavelength.

Because a quarter-wavelength line is frequently used as a linear transformer, it is convenient to calculate the length of a quarter-wave line directly. The formula is

$$\textit{Length} \text{ (feet)} = \frac{246}{f} \cdot V \qquad \textbf{(10-G)}$$

where the symbols have the same meaning as above.

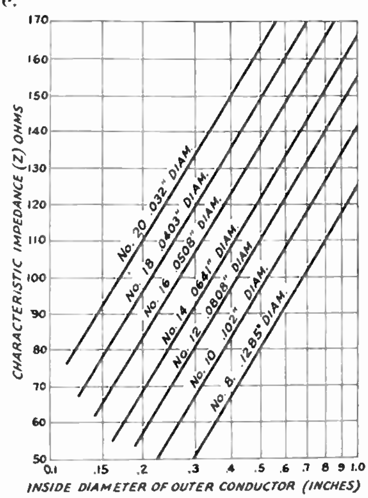


Fig. 10-11 — Chart showing characteristic impedance obtained with various air-insulated concentric lines.

Losses in Transmission Lines

There are three ways by which power may be lost in a transmission line: by radiation, by heating of the conductors (I^2R loss), and by heating of the dielectric, if any. Loss by radiation will occur if the line is unbalanced and, particularly with open-wire lines, may greatly exceed the heat losses. It can be reduced to a minimum by terminating the line in a balanced load and by symmetrical construction.

Heat losses in both the conductor and the dielectric increase with frequency. Conductor losses also are greater the lower the characteristic impedance of the line, because a higher current flows in a low-impedance line for a given power input. The converse is true of dielectric losses because these increase with the voltage, which is greater on high-impedance lines. The dielectric loss in air-insulated lines is negligible (the only loss is in the insulating spacers) and such lines operate at high efficiency when radiation losses are low. In solid-dielectric lines most of the loss is in the dielectric, the conductor losses being small.

It is convenient to express the loss in a transmission line in decibels per unit length, since the loss in db. is directly proportional to the line length. Losses in various types of lines operated without standing waves (that is, terminated in a resistive load equal to the characteristic impedance of the line) are given in Table 10-I. In these figures the radiation loss is assumed to be negligible.

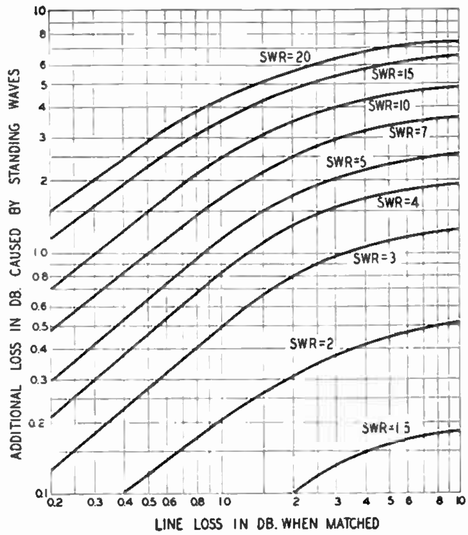


Fig. 10-12 — Effect of standing-wave ratio on line loss. The ordinates give the *additional* loss in decibels for the total line loss, under perfectly-matched conditions, shown on the horizontal scale.

When there are standing waves on the line the power loss increases as shown in Fig. 10-12. Whether or not the increase in loss is serious depends on what the original loss would have been if the line were perfectly matched. If the loss with perfect matching is very low, a large s.w.r. will not greatly affect the *efficiency* of the line — i.e., the ratio of the power delivered to the load to the power put into the line.

Example: A 150-foot length of RG-11/U cable is operating at 7 Mc. with a 5-to-1 s.w.r. If perfectly matched, the loss from Table 10-I would be 1.5 × 0.41 = 0.615 db. From Fig. 10-12 the additional loss because of the s.w.r. is 0.73 db. The total loss is therefore 0.615 + 0.73 = 1.345 db. The total power loss is just sufficient to make a detectable change in signal strength when observing conditions are ideal, but the additional loss caused by the s.w.r. is below the detectable (1 db.) level. With perfect matching the line efficiency is approximately 87 per cent. With the 5-to-1 s.w.r. the efficiency drops to about 73.5 per cent.

An appreciable s.w.r. on a solid-dielectric line may result in excessive loss of power at the higher frequencies. Such lines, whether of the parallel-conductor or coaxial type, should be operated as nearly flat as possible, particularly when the line length is more than 50 feet or so. As shown by Fig. 10-12, the increase in line loss is not too serious so long as the s.w.r. is below 2 to 1, but increases rapidly when the s.w.r. rises above 3 to 1. Tuned transmission lines such as are used with multiband antennas always should be air-insulated, in the interests of highest efficiency.

TABLE 10-I
Transmission-Line Velocity Factors and Attenuation

Type of Line	*Velocity Factor V*	*Attenuation, db./100 ft.; Mc.: for line terminated in Zo.*						*Capacitance per foot µµfd.*
		3.5	*7*	*14*	*28*	*50*	*144*	
Open-wire, 400 to 600 ohms [1]	0.975	0.03	0.05	0.07	0.1	0.13	0.25	
Parallel-tubing [1, 2]	0.95							
Coaxial, air-insulated [1]	0.85	0.2	0.28	0.42	0.55	0.7	1.4	
RG-8/U (53 ohms)	0.66	0.28	0.42	0.61	1.0	1.4	2.6	29.5
RG-58/U (53 ohms)	0.66	0.53	0.8	1.2	1.9	2.7	5.1	28.5
RG-11/U (75 ohms)	0.66	0.27	0.41	0.61	0.92	1.3	2.4	20.5
RG-59/U (73 ohms)	0.66	0.56	0.82	1.4	1.8	2.5	4.6	21.0
Twin-Lead, 300 ohms [3]	0.82	0.19	0.28	0.42	0.61	0.85	1.5	5.8
Twin-Lead, 150 ohms	0.77	0.24	0.35	0.52	0.76	1.05	1.9	10
Twin-Lead, 75 ohms	0.68	0.37	0.64	1.1	1.9	3.0	6.8	19
Transmitting Twin-Lead, 300 ohms [4]	0.84	0.16	0.23	0.34	0.5	0.69	1.24	
Transmitting Twin-Lead, 75 ohms	0.71	0.29	0.49	0.82	1.4	2.1	4.8	
Rubber-insulated twisted-pair or coaxial [5]	0.56 to 0.65	0.96	1.6	2.5	4.2	6.2	13	

[1] Average figures for air-insulated lines taking into account effect of insulating spacers.
[2] Losses between open-wire line and air-insulated coaxial cable. Actual loss with both open-wire and parallel-tubing lines is higher than listed because of radiation, especially at higher frequencies.
[3] Data for Amphenol types 14-056 and 14-271.
[4] Data for Amphenol type 14-076.
[5] Approximate figures for good-quality rubber insulation.

Unbalance in Parallel-Conductor Lines

When installing parallel-conductor lines care should be taken to avoid introducing electrical unbalance into the system. If for some reason the current in one conductor is higher than in the other, or if the currents in the two wires are not exactly out of phase with each other, the electromagnetic fields will not cancel completely and a considerable amount of power may be radiated by the line.

Maintaining good line balance requires, first of all, a balanced load at its end. For this reason the antenna should be fed, whenever possible, at a point where each conductor "sees" exactly the same thing. Usually this means that the antenna system should be fed at its electrical center. Even though the antenna appears to be symmetrical, physically, it can be unbalanced electrically if the part connected to one of the line conductors is inadvertently coupled to something (such as house wiring or a metal pole or roof) that is not duplicated on the other part of the antenna. Every effort should be made to keep the antenna as far as possible from other wiring or sizable metallic objects. The transmission line itself will cause some unbalance if it is not brought away from the antenna at right angles to it for a distance of at least a quarter wavelength.

In installing the line conductors take care to see that they are kept away from metal. The minimum separation between either conductor and all other wiring should be at least four or five times the conductor spacing. The shunt capacitance introduced by close proximity to metallic objects can drain off enough current (to ground) to unbalance the line currents, resulting in increased radiation. A shunt capacitance of this sort also constitutes a reactive load on the line, causing an impedance "bump" that will prevent making the line actually flat.

Coupling the Transmitter to the Line

In very general terms, the problem of coupling the transmission line and transmitter together is one of transforming the input impedance of the line into a value of impedance that will "load" the transmitter properly — that is, cause it to deliver the desired power output at as high efficiency as the transmitter design will permit. This is a question of impedance matching, and the impedance that must be matched is the value of resistance into which the tubes in the final stage of the transmitter should work. The value of this resistance is determined by the choice of tube operating conditions. The tubes are working into the proper resistance when the final tank circuit is tuned to resonance and the loading is such that the tubes are drawing rated plate current, as described in Chapter Six. The proper value of load resistance is thus reached automatically when the coupling is adjusted to bring the plate current up to the normal operating value. It is therefore not at all necessary to know what value of resistance is required. It is sufficient to note that, in general, it is in the neighborhood of a few thousand ohms, and is higher the higher the plate-voltage/plate-current ratio of the final stage.

The input impedance of the line can assume a wide range of values. As described earlier, it may be very much higher or very much lower than the impedance of the load at the end of the line, unless the line is matched to the load. Furthermore, it may or may not be a pure resistance, depending on the s.w.r., the line length, and the characteristics of the load.

Transforming Impedances

It was explained in Chapter Two that a resistive load tapped across part of a tuned circuit is equivalent to a higher value of resistance connected in parallel with the whole circuit. In other words, there is a transformer action in such an arrangement that enables us to change the value of a given resistance, such as R in Fig. 10-13A, into a new and higher value when the source of power looks into the terminals AB. Given reasonable values for L and C, the resistance looking into AB is determined practically wholly by the value of R and the position of the tap, so long as LC is tuned to resonance with the applied frequency. This is because the resonant impedance of LC alone (with R disconnected) is usually very high

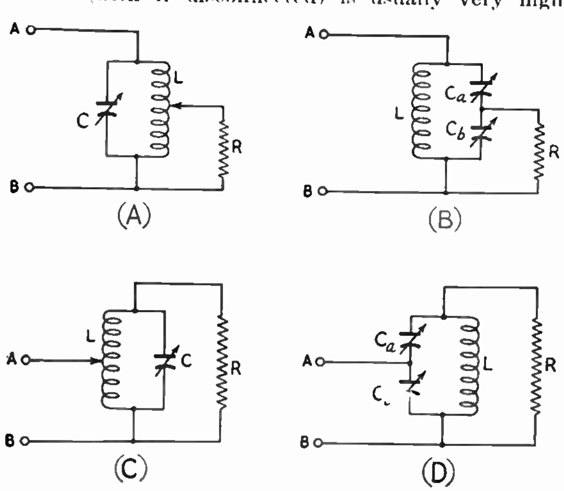


Fig. 10-13 — Using a resonant circuit for matching impedances.

compared with the resistance, R, of any practical load likely to be used, and also compared with any resistance that might be required between the terminals AB.

Fig. 10-13B shows a circuit that also provides a method for impedance transformation, using a capacitance voltage divider instead of tapping on the inductance. In this case, decreasing the capacitance of C_b (while increasing the capacitance of C_a correspondingly to maintain resonance) has the same effect as moving the tap toward the top of the coil in Fig. 10-13A. This type of circuit gives very smooth control. However, variable condensers of impracticable size would be necessary, to give as wide a range of impedance transformation as the circuit at A.

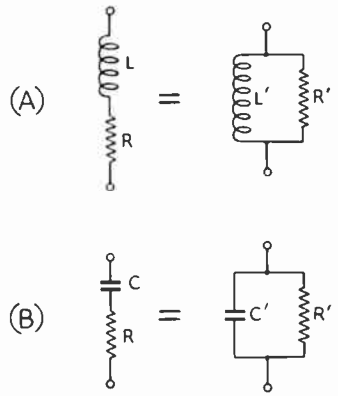


Fig. 10-14 — Series and parallel equivalents of a line whose input impedance has both reactive and resistive components.

When an r.f. amplifier is coupled to a transmission line the line impedance very seldom is larger than the load impedance required by the amplifier. However, should such a case arise the same circuits can be used by reversing the terminals. This is shown at C and D in Fig. 10-13. With R connected across the whole circuit, its resistance can be transformed to a lower value when the input terminals are tapped across part of the coil, as at A, or across C_b in Fig. 10-13B. The nearer the tap is to the bottom end of the coil, or the larger the capacitance of C_b compared with C_a, the smaller the resistance between terminals AB.

Complex Loads

In the foregoing it was assumed that the load, R, was a pure resistance. However, the input impedance of a line is more likely than not to have a reactive as well as a resistive component. This means, basically, that the current flowing into the line is not in phase with the voltage applied to the line. To represent such a condition by circuit symbols we can assume the input impedance of the line to consist either of a reactance (coil or condenser) in series with a resistance, or a reactance in parallel with a resistance. It does not matter which we choose, so long as the values assigned to the resistance and reactance are such that if the voltage were applied to the circuit instead of to the line, the current that flows would have exactly the same amplitude and phase angle as it actually does at the input terminals of the line.

These equivalent circuits are shown in Fig. 10-14. In practical work with lines it is not necessary to know the values of R, L or C. It is sufficient to know that they *symbolize* a condition that exists at the input end of the line and then to know what to do about them. A few general points are worth noting: Given a fixed value of voltage, if the current at the input end of the line is high, then the impedance is relatively low; if the current is low, the impedance is relatively high. If the current is very nearly in phase with the voltage the reactance in the *series* equivalent circuit is small, but the reactance in the *parallel* equivalent circuit is large. On the other hand, if there is a considerable phase difference between current and voltage the reactance is large in the equivalent series circuit and is low in the equivalent parallel circuit. (In visualizing these reactances as coils and condensers it must be remembered that "large" and "small" are relative terms; for example, a "large" inductance at 28 Mc. would be a "small" inductance at 3.5 Mc. Also, the larger the capacitance of a condenser the smaller its reactance.)

Now suppose that a reactive line is to be connected to our impedance-transforming resonant circuit. Let us choose the parallel equivalent circuit, since it is somewhat easier to picture what happens. Fig. 10-15A shows a load with inductive reactance tapped across part of the resonant circuit (corresponding to Fig. 10-13A), and a load with capacitive reactance is shown in Fig. 10-15B. Imagine for the moment that the load has only reactance; the resistive component, R, is disconnected. Then, just as in the pure-resistance case previously

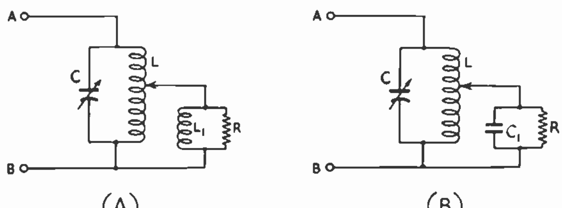


Fig. 10-15 — Circuit equivalent of a reactive line connected to a resonant circuit for impedance matching.

discussed, a small reactance tapped across the coil L will appear as a larger reactance across the whole circuit, or between the input terminals AB. Thus, connecting a coil, L_1, across part of L is equivalent to connecting a larger coil across the whole circuit. Connecting a condenser, C_1, across part of L is equivalent to connecting a *smaller* condenser (larger reactance) across the whole circuit.

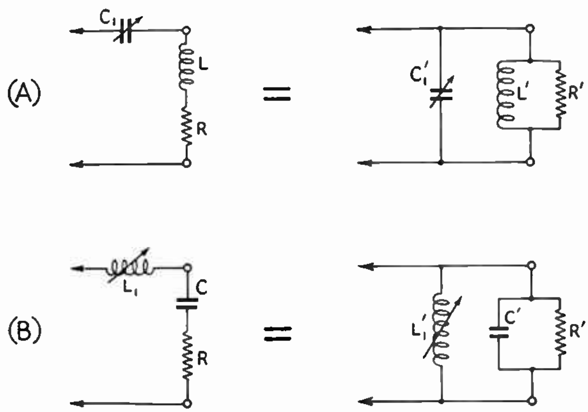


Fig. 10-16 — Methods for canceling the reactive component of the input impedance of a transmission line. In A the line input impedance is represented by L and R in series, or by L' and R' in parallel, and in B by C and R in series, or by C' and R' in parallel.

In either case this equivalent shunting reactance detunes the LC circuit from resonance, and C must be readjusted to bring it back. In the case of Fig. 10-15A, the capacitance of C must be increased because the "reflected" reactance in parallel with L decreases the total inductance (inductances in parallel) and so tunes the circuit to a higher frequency. The opposite is the case in Fig. 10-15B; the shunting reactance is capacitive and increases the total capacitance. Consequently the capacitance of C must be decreased to bring the circuit back to resonance.

The over-all effect, then, of coupling a reactive load to the circuit is to cause detuning as well as to cause the desired resistance loading. If the reflected reactance is large, corresponding to connecting a very large coil or a very small condenser across the whole LC circuit, it is readily possible to retune the circuit to resonance by adjusting C. The nearer the tap to the top end of L, the greater the change required in the tuning. But this simple method of compensating for the reactive component of the load is not always sufficient. In some cases the tap has to be moved so far up the coil, in order to obtain the right value of resistance loading, that the tuning condenser, C, no longer has sufficient range to compensate for the reflected reactance. When such a condition exists it is difficult, and sometimes impossible, to couple the desired amount of power to the transmission line.

Canceling Line Reactance

The remedy for this condition is to make the input end of the line look like a pure resistance *before* it is tapped on the impedance-transforming circuit. This can be done by "tuning out" the reactance of the line, by inserting a reactance of the same value but of the opposite kind. Again we have our choice between considering the line to be represented by reactance and resistance in series, or by reactance and resistance in parallel. The circuits are shown in Fig. 10-16. In A, a condenser, C_1, is used to cancel out the inductive reactance of the line, and in B an inductance, L_1, is used to cancel capacitive reactance. The same value of capacitance cannot be used for C_1 and C_1' under a given set of conditions because, as explained earlier, L and L' do not have the same values. For example, if L is small its parallel equivalent, L', is large, so a large capacitance would be required at C_1 and a small capacitance at C_1'. Because of limitations in practicable components (particularly in the capacitance range of variable condensers), there are conditions where the series circuit is the easiest to set up, from a practical standpoint. In others, the parallel circuit is easier to get working. For the large majority of cases either circuit will work equally well; from the standpoint of convenience, the parallel circuit is probably better.

To summarize, then, we have three general cases as shown in Fig. 10-17. If the line is purely resistive, or so nearly so that such reactance as is reflected across the LC circuit can be tuned out by readjusting C, the circuit at A may be used. Where the line shows more pronounced reactive effects, the line reactance can be tuned out, as indicated at B and C, so that the load tapped on L is purely resistive. It is easy to tell which should be used, inductance or capacitance, to compensate for the line reactance. If the line only (Fig. 10-17A)

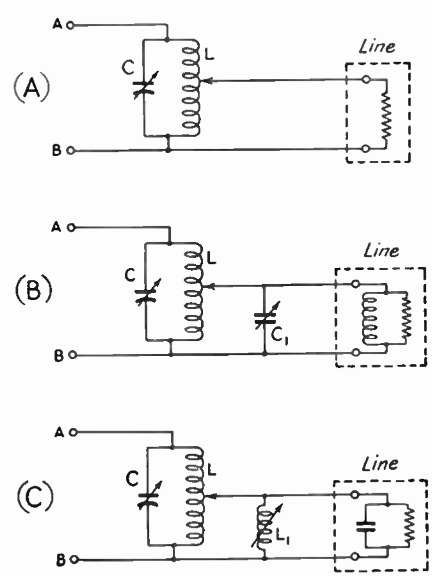


Fig. 10-17 — Methods of canceling line input reactance combined with impedance transformation.

is tapped across a very small portion of L, C will have to be readjusted slightly to bring the LC circuit back to resonance. If the capacitance of C has to be increased, a condenser, C_1, should be connected across the input terminals of the line. If the capacitance of C has to be decreased, an inductance, L_1, should be connected across the line. In either case the compensating reactance, C_1 or L_1, should be adjusted in value until the setting of C, for resonance with the applied frequency, is the same whether or not the line is tapped on L. When this condition is reached the loading may be adjusted by changing the tap position until the amplifier takes the desired plate current.

PRACTICAL COUPLING SYSTEMS

In practical work the two primary functions that a coupling system must perform — tuning out the line reactance, if any, and providing a method for control of loading on the transmitter — are not always enough. For one thing, it is desirable that the coupling system be such that the transmission line will operate only in the way it is intended that it should. For another, the coupling system should prevent transfer of any of the harmonic energy that always is present in the output of a transmitting amplifier. Both these points will be considered later in this section. For the moment, let us take a look at some of the simpler coupling systems.

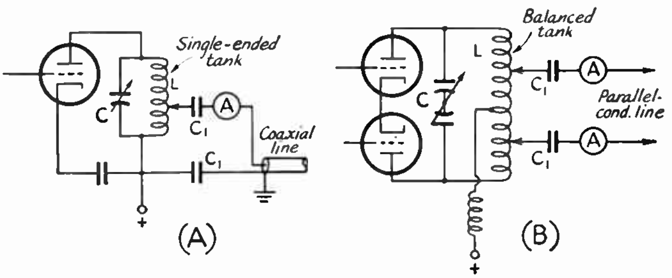


Fig. 10-18 — Simple methods of coupling to a transmission line. The blocking condensers, C_1, should be 0.001-μfd. (or larger) mica condensers having a voltage rating in excess of the maximum d.c. voltage applied to the final amplifier (including the voltage applied on modulation up-peaks). The coaxial line can be coupled to a balanced tank circuit by connecting the grounded shield to the center of the coil (through a blocking condenser) and tapping the inner conductor on one side of the center. The parallel-conductor line requires a balanced tank circuit.

The possibility of tapping the input end of the transmission line directly on the final-amplifier tank suggests itself from the discussion earlier. This method will work when the input impedance of the line is purely resistive, or nearly so. It can therefore be used with nonresonant or untuned lines, or with a resonant line when the line has the right length. As explained earlier, the input impedance of the line will be resistive when its length is a multiple of a quarter wavelength, provided the load at the output end of the line is a pure resistance. This will be so if the antenna itself is resonant, but will not be true if the antenna length is not correct for the operating frequency. The circuits are shown in Fig. 10-18. If the final amplifier is series-fed so that the tank circuit is "hot" with the plate voltage, it is necessary to connect a blocking condenser between the tank and the line. These circuits, although simple, are not recommended except perhaps in emergencies; there is little or no discrimination against harmonic frequencies.

Adjustment of this type of coupling is simple. First, resonate the amplifier tank circuit, with the line disconnected, by setting the tank condenser, C, to the minimum plate current point. Then tap the line across a turn or two of the tank coil, and readjust C for minimum plate current. The new minimum will be higher than with no load on the tank. Continue increasing the number of turns between the line taps, readjusting C each time, until the minimum plate current is the desired full-load value.

R.F. Ammeters

The r.f. ammeters shown in Fig. 10-18 and subsequent coupling circuits are useful accessories. The input impedance of the line is unaffected by any adjustments made in the coupling system (except for the effects of stray capacitance, as discussed later) so the greater the current flowing into the line the larger the amount of power delivered to the load. Measurement of r.f. current thus gives a check on the adjustment procedure and indicates when the largest power output is being obtained. Obviously, an adjustment that increases the input to the final stage of the transmitter without causing the line current to increase has simply increased the losses without increasing the output.

In the case of parallel-conductor lines two ammeters are shown, one in each conductor. This gives a check on line balance, since the two currents should be the same. It is not actually necessary to use two instruments; one ammeter can be switched from one side of the line to the other for comparative measurements. Also, it is to be understood that any current-indicating device (such as a flashlight lamp) that will work at r.f. may be used as a substitute for an actual ammeter.

The scale range required depends on the input impedance of the line and the power. The current to be expected can readily be calculated from Ohm's Law when the line is flat. In other cases the s.w.r. and the length of the line must be considered. The maximum current will occur when there is a current loop at the input end of the line, and if the load impedance and line impedance are known the input impedance at a current loop can be calculated from the formulas given earlier.

The ammeters are less useful when the input impedance of the line is high, because in that case the input current is quite small. It is to be noted that the value of current does not indicate, in any absolute sense, how well the system as a whole is working unless the actual value of the resistance component of the line input impedance is known. Current measurements taken on different lines, or on the same line if its length in wavelengths is changed, are not directly comparable.

Inductive Coupling

The circuits shown in Fig. 10-19, like those in Fig. 10-18, are useful only with lines having purely-resistive input impedance. The pick-up coil, which is inductively coupled to the tank coil, is in fact simply a substitute for the tapped portion of the tank coil in Fig. 10-18. The number of turns required in the pick-up coil depends upon the resistance represented by the input end of the line. For flat lines, the number is governed by the characteristic impedance of the line. For 50- or 70-ohm lines it may range from one or two turns, at frequencies of the order of 14 to 28 Mc., to several turns at 3.5 Mc. For higher-impedance lines it may take half as many turns as there are in the tank coil, to get adequate coupling. In both cases the coupling between the coils will have to be very tight. The link windings provided on commercial coils are not usually adequate for this type of coupling except for low-impedance lines at the higher frequencies. When the number of turns on the pick-up coil is fixed, the loading on the final amplifier can be varied by varying the coupling between the two coils. Inductive coupling of this type is somewhat better than direct coupling from the standpoint of harmonic transfer.

Pick-up coil coupling introduces some reactance into the tank circuit, because of the leakage reactance of the coupling coil. This must be compensated for by retuning the final tank circuit when the desired degree of coupling is reached. If very much retuning is required, or if the amplifier loads with loose coupling between the two coils, it is an excellent indication that the line is not actually flat.

When a "swinging-link" assembly is used to obtain this type of coupling, the loading on the final amplifier can be adjusted to the desired value by varying the coupling between the two coils. The tank condenser, C, should be readjusted to minimum plate current each time the coupling is changed. If the desired loading cannot be obtained there is no alternative but to use a different coupling system.

The pick-up coil may be wound directly over the final tank coil, in which case the correct number of turns may be determined by

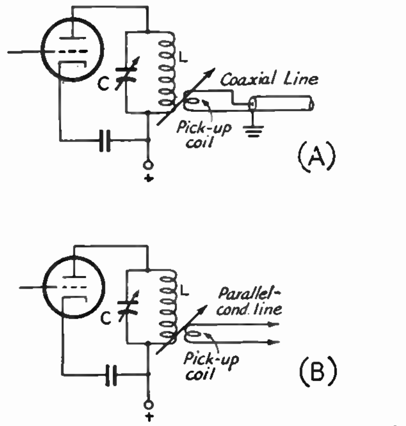


Fig. 10-19 — Using an untuned pick-up coil to couple to a transmission line. The method of adjustment is discussed in the text.

trial. The insulation between the coils must be adequate for the plate voltage used, if the amplifier is series-fed.

Series and Parallel Tuning

The circuits shown in Fig. 10-20 are useful with parallel-conductor lines operating at a relatively-high standing-wave ratio, particularly when the line length is such as to make the input impedance substantially a pure resistance. Assuming that the antenna is resonant, the optimum line lengths will be multiples of a quarter wavelength at the operating frequency. When the s.w.r. is high, the impedance at such points is considerably higher or considerably lower than the characteristic impedance of the line.

In these circuits the secondary, consisting of L_1, C_1 (and C_2, in the series circuit) and the input impedance of the line, is tuned to the operating frequency. As explained in Chapter Two, the degree of coupling between two resonant circuits is determined by their Qs, and it is necessary to keep the Qs fairly high (of the order of 10 or so). Assuming that the input impedance of the line is purely resistive, it can be inserted in series with the circuit (as in A) if its value is below about 100 ohms. The Q of the secondary circuit then can be brought to the proper value by making the reactance of L_1 of the order of 500 to 1000 ohms and setting the total capacitance of C_1 and C_2 to tune the circuit to resonance. With this type of tuning the current flowing into the line is rather large; in other words, the system is suitable for coupling into the line at a current loop.

On the other hand, if the line impedance is of the order of a few thousand ohms or more — which it will be at a voltage loop when the s.w.r. is high — the secondary circuit cannot be made to take power from the transmitter if the line resistance is inserted in series. The Q of the secondary circuit would be far too low to give adequate coupling. In such a case the parallel-tuned circuit at B may be used. As exmitter. To take care of cases where the input impedance of the line has a considerable reactive component, provision is made for switching in either a shunt capacitance or inductance, both of which are variable (see earlier discussion). The coupling should be variable at least at one end of the link circuit.

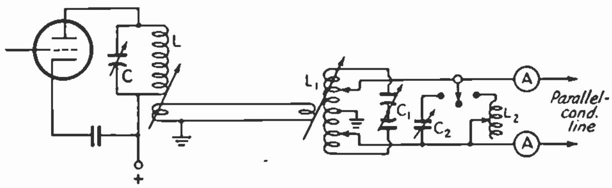


Fig. 10-22 — "Universal" antenna-coupling system. This circuit can be used with both resonant and nonresonant parallel-conductor lines.

In general, it is advisable to make the inductance of L_1 about the same as that of L, and to use for C_1 a condenser of the same capacitance as that used for C. The voltage rating for C_1 also should be the same as that of C. In other words, L_1C_1 may be a duplicate of LC for the operating frequency in use. The link coils can consist of two or three turns at each end. Provision should be made for tapping L_1 at frequent intervals — every turn, if possible. C_2 should have as large a maximum capacitance as is convenient — 250 to 500 $\mu\mu$fd. — but its voltage rating need not be high in the average case. For most installations where the power output does not exceed a few hundred watts a plate spacing of the order of 0.025 to 0.05 inch is sufficient. The inductance L_2 can consist of 20 or 25 turns approximately 2 inches in diameter and spaced 8 to 10 turns to the inch. The coil should be tapped every few turns.

The tuning procedure is as follows: First, disconnect the feeder taps on L_1 and use the loosest possible coupling, through the variable link coupling, to the final tank circuit. Tune C_1 until the plate current rises to a peak, indicating that L_1C_1 is resonated, and note the setting of C_1. Cut C_2 and L_2 out of the circuit and then connect the line taps across a turn or two at the center of L_1. Readjust C_1 to resonance, as indicated by a rise in plate current. It should be necessary to use closer coupling to get an observable change in plate current with the line connected. Note the new setting of C_1. If the capacitance is lower, switch in L_2 and find the tap that permits returning C_1 as nearly as possible to its original setting; if the capacitance is higher, switch in C_2 and adjust it to bring C_1 back to the original setting. Then increase the coupling, keeping C_1 at resonance as indicated by maximum plate current, and keeping C at resonance as indicated by minimum plate current. Continue until the minimum plate current reaches the desired load value. If C_1 flashes over as the coupling is increased, or if tuning C_1 back and forth a small amount either side of resonance makes it necessary to change the setting of C appreciably to maintain the final tank in resonance, the taps on L_1 are too close together. Move each tap one turn toward the ends of L_1, and again try increasing the coupling for rated load on the amplifier. When the proper loading is obtained, the tuning of L_1C_1 will be reasonably sharp, and changing the coupling will not necessitate more than "touching up" C to maintain resonance. If the taps on L_1 are too far apart the antenna tank circuit, L_1C_1, will be loaded heavily and its tuning will be broad. Under these conditions it may also be impossible to load the amplifier to rated plate current, even with the tightest available coupling. On the other hand, if the taps on L_1 are too close together the antenna tank will be too lightly loaded; its tuning will be critical and will affect the tuning of the plate tank circuit to a marked degree, and L_1 may overheat when the coupling is adjusted to make the amplifier take normal input.

When the reactive effects at the input end of the line are small, neither C_2 nor L_2 will be required. When this is the case, the setting of C_1 for resonance will not change much when the line is tapped on L_1. The greater the number of turns between the taps, the greater the detuning of the antenna tank by a given amount of reactance in the transmission-line input impedance.

This coupling system is equally effective with flat lines or those operating at a high s.w.r. If the line is actually flat, C_2 and L_2 will not be needed and the resonance setting of C_1 will not be affected by connecting the line. Regardless of the s.w.r., the positions of the line taps will depend on the resistive component of the line input impedance. If the resistance is low, the taps will be close together; if it is very high, the taps may have to be set right at the ends of L_1.

Coupling to Coaxial Lines

The principles of coupling to coaxial lines are just the same as for coupling to parallel-conductor lines. However, this type of line is unbalanced to ground, has inherently low impedance, and always should be operated with a low standing-wave ratio. The input impedance of a properly-operated coaxial transmission line therefore will be principally resistive, and of a value varying between perhaps 30 to 100 ohms, depending on the type of line and the s.w.r.

It is possible to couple such a line by means of a small coil inductively coupled to the final tank coil, as shown in Fig. 10-19A. The small amount of reactance introduced by the pick-up coil — and by the line, if the s.w.r. is slightly greater than 1 — can readily be tuned out by adjustment of the final tank condenser. However, additional selectivity is desirable for the purpose of reducing harmonic transfer from the final tank. Circuits are shown in Fig. 10-23. Except that it is adapted for single-ended rather than balanced operation, the circuit at A operates in much the same way as the circuit in Fig. 10-22. Also, because the load is known to be in the region of 100 ohms or less, it is possible to tap it across a capacity voltage divider (see earlier discussion) for impedance matching. This avoids the necessity for tapping L_1.

The circuit of Fig. 10-23B is similar in operation to that at A, but dispenses with the link circuit. For convenience, it uses a link coil on the final tank for inductive transfer of energy, the rest of the inductance in the antenna tank circuit being made up by L_1.

In the circuit at A, L_1 may be the same as L; in B, L_1 plus the pick-up coil should have about the same inductance as L. Except at perhaps 28 Mc., it is satisfactory, practically, to make L_1 the same as L in this circuit also, since the pick-up coil will not ordinarily have much inductance itself. In both circuits C_2 should have about the same capacitance as C, and C_1 should have approximately the value suggested in Fig. 10-23.

To adjust the circuit, set C_1 at maximum, loosen the coupling between L and the link or pick-up coil, and tune C_2 to resonance. This will be indicated, as usual, by a rise in the amplifier plate current. Adjust C to minimum plate current and increase the coupling in small steps, reresonating C_2 and C each time, until the amplifier plate current is normal. The loading on the antenna tank circuit is least when C_1 is at maximum capacitance, and increases when the capacitance of C_1 is decreased (with C_2 increased correspondingly to maintain resonance). The symptoms of under- and over-loading of the antenna tank are the

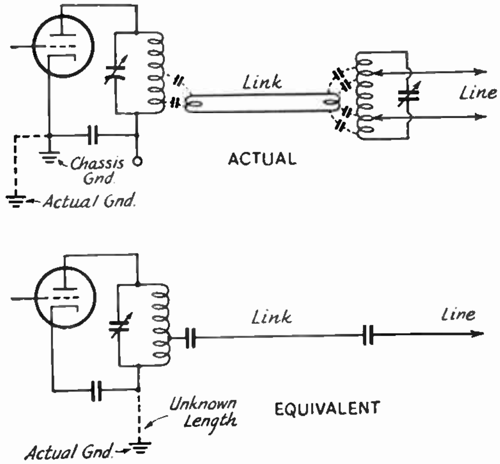


Fig. 10-24 — The stray capacitive coupling between coils in the upper circuit leads to the equivalent circuit shown below. The effect on the performance of the antenna system is discussed in the text.

same as described in connection with the universal antenna coupler. Adjust the loading by means of C_1, so that at normal plate input the antenna tank tuning is reasonably sharp and the setting of C is not greatly affected when C_2 is tuned a small amount either side of resonance.

Stray Coupling

In most of the circuits in Figs. 10-18 to 10-23, inclusive, a single-ended tank circuit has been indicated for the final amplifier. The amplifier itself has been shown only sketchily. The fact is that any type of antenna coupling circuit can be used with any type of amplifier circuit — screen grid or neutralized triode, single-ended or push-pull. However, the actual arrangement, physically, of the circuit elements usually has an important bearing on the performance of the system. As it happens, a coupling system that is poorly designed, constructionally speaking, usually will do what it is supposed to do. But, equally important, it may do a lot of things it is *not* supposed to do.

Most of the unwanted effects that occur on transmission lines can be traced to stray capacitances in the system. Fig. 10-24 is an illustration. The upper drawing shows the ordinary link-coupled system as it might be used to couple into a parallel-conductor line. Inasmuch as a coil is a sizable metallic object, it will have capacitance to any other metallic objects in its vicinity, including other coils. Consequently there is capacitance between the final tank coil and its associated link coil, and between the antenna tank coil and its link. These capacitances

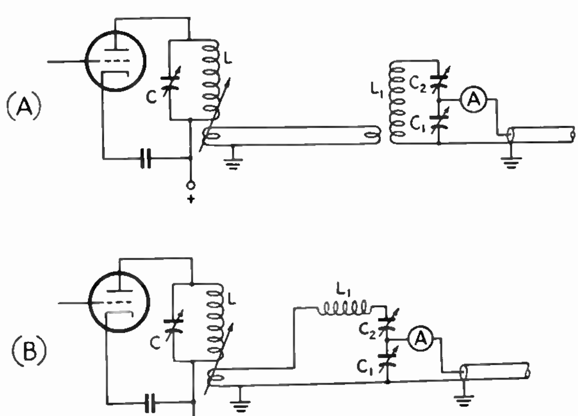


Fig. 10-23 — Coupling to coaxial lines. These circuits are used for harmonic suppression when working into a nonresonant coaxial line. Recommended capacitance values for C_1 are as follows: 28 Mc., 100 $\mu\mu$fd.; 14 Mc., 200 $\mu\mu$fd.; 7 Mc., 400 $\mu\mu$fd.; 3.5 Mc., 800 $\mu\mu$fd.

are small, but not negligible. In addition, the transmitter, particularly with metal-chassis construction, has appreciable capacitance to ground. Even if it did not, there is always a path from the transmitter to ground through the power wiring and the many stray capacitances associated with it.

There is a fundamental difference between the inductive coupling between coils and the capacitive coupling between them. Inductive coupling induces a voltage in the secondary coil that causes a current to flow, in common terminology, "around" the circuit. In Fig. 10-24, this means that the same current flows in both conductors of the link but, if the wires are parallel, the current flows in opposite directions in the two as it completes its travel around the loop. The same is true of the currents in the two conductors of the line. But with stray capacitive coupling the voltages at all points on the secondary coil are essentially in phase; for this type of coupling the secondary coil is just a mass of metal. Consequently, whatever current flows in the link (or in the line) flows in the *same* direction in both wires. Although both the link and line have two conductors and apparently form an ordinary go-and-return circuit, to the currents that flow as a result of capacitive coupling they simply look like a pair of conductors in parallel — in effect, that is, like a single conductor. The equivalent circuit is shown in the lower drawing in Fig. 10-24.

This single-wire circuit is an antenna system in itself, working in conjunction with a ground lead of unknown composition and length. It includes the regular antenna as well as the entire transmission line. If the various lengths happen to be just right, a fairly-large "parallel" current of this type can flow in it. This means that a considerable proportion of the total power output of the transmitter can be wasted in losses and radiation from a very undesirable sort of antenna system. Furthermore, despite the tuned tank circuits in the amplifier and antenna coupler, harmonic currents will flow in such an "antenna" even more readily than the fundamental current.

There are other undesirable results, too. The fact that the power wiring becomes part of an "antenna" system means that the transmitter itself may perforce be at a considerable r.f. potential above ground. The chassis becomes "hot" with r.f., r.f. feed-back is prone to occur in speech equipment, and a considerable amount of r.f. power may be pumped into receiving and other equipment connected to the same a.c. power outlet. (A similar type of coupling in the input circuits of a receiver leads to stray pick-up of signals that may partially or completely mask the directive effects of the proper antenna.) On top of all this, it is impossible to tell much about the operation of the transmission line because the parallel current is more or less in phase with the regular line current in one wire and out of phase with it in the other. Thus the resultant currents in the two wires are unbalanced, and there is no way to separate the "parallel" and "line" currents in measurement.

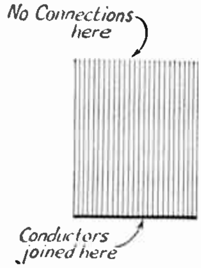


Fig. 10-26 — **The Faraday screen. Stiff wire or small-diameter rod may be used for the conductors, separated by a distance about equal to the diameter of the wire or rod. The dimensions of the screen should be greater than the diameters of the coils to be shielded from each other.**

These effects can only be eliminated if the stray capacitances are eliminated. However, they can be reduced by arranging the coils so the amount of energy coupled from the primary to the secondary is small, even though the capacitance itself still exists. This can be done by using a link coil that is physically small — that is, has few turns — and coupling it to the "cold" point on the tank coil. The cold point will be at the end of the coil that is grounded for r.f., either directly or through a by-pass condenser, in the case of single-ended tanks. In balanced tank circuits, the cold point is at the center. The coupling is further reduced if one side of the link circuit is grounded to the transmitter chassis as close as possible to the point where the tank itself is grounded. If the link is at the end of the tank coil the side farthest from the tank should be grounded, as indicated in Fig. 10-25A. If the link is wound *over* one end of the tank coil, ground the side toward the hot end of the tank, as indicated in Fig. 10-25B. With a balanced tank circuit the link should be at the center of the coil. In this case the best point to ground is the center of the link coil, but if this is impracticable good results will be secured by grounding either end of the coil. Ground directly to the chassis and keep the lead as short as possible.

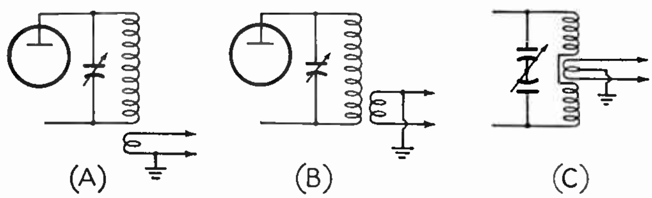


Fig. 10-25 — **Methods of coupling and grounding link circuits to reduce energy transfer through stray capacitance.**

This treatment of link circuits does not eliminate capacitive coupling. It simply makes it less troublesome, by making certain that the coupling occurs between parts of circuits that

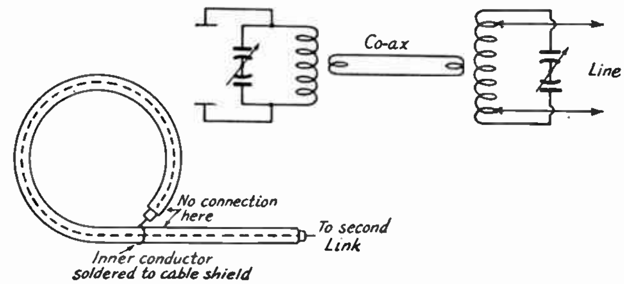


Fig. 10-28 — A shielded link coil constructed from coaxial cable. The smaller sizes of cable such as RG-59/U are most convenient, except when the coils have a diameter of 3 inches or more. For larger coils, RG-8/U or RG-11/U can be used.

are not at high r.f. voltage. However, there are cases, particularly with balanced tank circuits, where the point on the tank coil that is cold for the fundamental frequency is hot at the even harmonics. This means that even though the transmitter and line behave properly on the fundamental frequency, harmonics still can be radiated at considerable intensity. The only way to be sure that these effects do not exist is to eliminate the stray capacitance entirely.

Capacitive coupling between coils can be eliminated by means of a **Faraday screen.** This is a shield that prevents the electric field from one coil from reaching the other, but which has no effect on the magnetic field. As shown in Fig. 10-26, it consists of a group of parallel conductors, insulated from each other, and connected together at one end only. This forms an effective shield for the electric field, but since the conductors are open-circuited the voltages induced in them by the magnetic field cannot cause any current to flow. (Such current flow is essential to magnetic shielding with nonmagnetic materials, as explained in Chapter Two.)

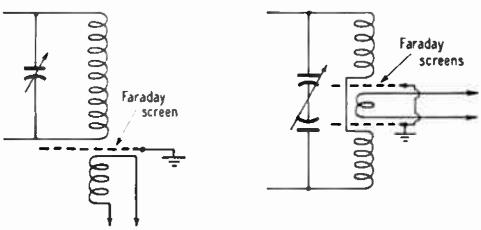


Fig. 10-27 — Installation of Faraday screens to eliminate capacitive coupling between coils.

The Faraday screen should be somewhat larger than the diameter of the coils with which it is used. It is simply mounted between the two coils that are to be shielded from each other, and then grounded to the chassis through a short lead, as indicated in Fig. 10-27. In the case of a balanced tank circuit with a swinging link, two shields must be used, one on each side of the link coil. In the case of fixed links wound over the tank coil, a satisfactory screen can be made by using several turns of the same type of coil, cutting them parallel to the axis to open-circuit the conductors, and then soldering them together at one end only. This shield can then be inserted between the tank coil and link, making sure that it is adequately insulated from both.

An alternative, and perhaps simpler, type of screening is shown in Fig. 10-28. In this case the inner conductor of a piece of coaxial cable is used to form a one-turn link. The outer conductor serves as an open-circuited shield around the turn, this shield being grounded to the chassis. The circuit to the link line is made by connecting the inner conductor to the outer conductor at the finish of the turn, as shown, and from there on the coaxial line is used to transfer the power to a second, and similar, link coil at the antenna tuner. This type of shielded link is simpler to make than the regular Faraday screen.

Aside from the adverse effects on the performance of the antenna system, stray capacitive coupling frequently is responsible for interference to near-by broadcast receivers. It is not difficult to appreciate that radiation taking place from transmission lines and power wiring is, in general, more likely to get into a broadcast receiver than radiation from an antenna that is intentionally kept away from other antennas — particularly when the receivers are connected to that same power wiring.

Harmonic Reduction

Besides its primary function of providing optimum power transfer from the transmitter to the antenna, the coupling system between the final stage and the transmission line should prevent harmonics from being transferred to the antenna along with the desired fundamental power. Harmonics that fall in the communication spectrum — i.e., up to about 30 Mc. — are usually suppressed to a satisfactory degree in a link-coupled antenna tuner of the type discussed in the preceding section. However, harmonics in the v.h.f. region can cause serious interference to television reception in the immediate vicinity of the transmitter, even though their amplitude is so low that they are not detectable at a distance.

As stated in Chapters Six and Nineteen, the reduction of harmonic radiation in television channels frequently involves more than preventing harmonics from being coupled from the final stage to the antenna system. Methods that have been found successful in preventing radiation from the transmitter itself are described in Chapter Six. They should be used to the extent necessary for preventing interference when the transmitter is working into a dummy antenna. Once it is determined that there is no interference on a dummy antenna, it is reasonably certain that if interference appears on reconnecting the regular antenna system, it is because harmonics are being coupled into the antenna.

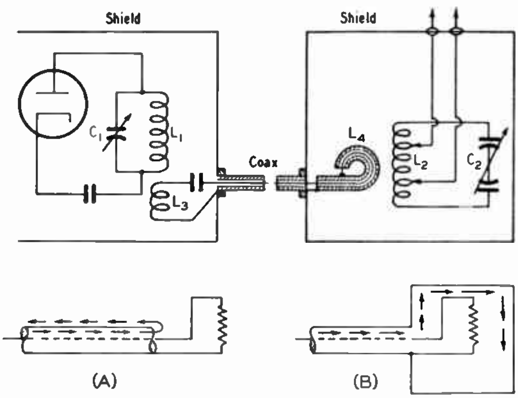


Fig. 10-29 — Recommended type of antenna coupler for reducing harmonic radiation. A ground on the rotor of C_2 may help in some cases; in others it may increase harmonic radiation. It should be tried both ways to see which gives the best results.

The link-coupled antenna coupler will do much to prevent harmonics from being transferred from the final stage to the transmission line. A shielded link or Faraday screen is highly desirable because it reduces harmonic transfer by stray capacitive coupling. Fig. 10-29 shows a suitable circuit arrangement for the most desirable form of coupler, one using coaxial cable for the link. Shielding around the transmitter and antenna coupler, along with coax fittings at both ends of the link line, are essential. Without them the harmonic currents can flow on the outside of the coax and will defeat the purpose of the system. It has been found possible to dispense with complete shielding of the antenna coupler if the circuits are mounted on a metal chassis so that the coax link can be terminated in a regular fitting, because the chassis tends to perform the same function as a shield in terminating the cable. As compared with connecting the transmission line directly to the output coil (L_3 in Fig. 10-29) measurements on this type of antenna coupler show a reduction of 25 to 30 db. in second-harmonic output from a 28-Mc. transmitter.

LOW-PASS FILTERS

Very great reduction of harmonic output can be secured by connecting a low-pass filter between the transmitter and the transmission line. Because stray coupling is hard to avoid with open-wire lines, such filters are most effective when used with coaxial lines. They can be used with any type of line and antenna system if inserted in a coax link between the final amplifier and an antenna coupler — for example, in the coax link in Fig. 10-29. By taking advantage of the so-called infinite-rejection points in *m*-derived filters (see Chapter Twenty-Four) very high attenuation of harmonics can be secured in particular television channels in which harmonics are most troublesome.

A simple filter of this type is shown in Fig. 10-31. This filter has two rejection frequencies and will give a minimum of 50 db. attenuation over any two selected channels in the 54–88 Mc. range. The attenuation in other channels varies from 20 to 40 db., depending on the frequency. In general, localities with a number of television stations fall into two groups. In one, the assignment pattern is Channels 2, 4 and 5 in the low band, and in the other Channels 3 and 6. The filter designs given in Fig. 10-31 are based on maximum attenuation in Channels 2 and 4 in the one case, and Channels 3 and 6 in the other. In either case the attenuation is ample for harmonics falling in the 174–216 Mc. range.

As shown in Fig. 10-32, the components are laid out in essentially the same form as in the circuit diagram. The condenser rotors are grounded to the aluminum plate on the side nearest the coax terminals, to keep the return paths as short as possible. The coils are mounted at right angles to reduce the coupling between them. A shield folded from a piece of aluminum is placed about the center condenser to reduce capacitive coupling between the three units. The other baffle shield similarly is used to reduce the coupling between L_1 and L_2.

The variable condensers are best adjusted by setting them to obtain maximum harmonic suppression while observing the interference in a television picture. C_1 and C_2 are both adjusted to the lower of the two channels, and C_3 to the higher of the two. If such a test

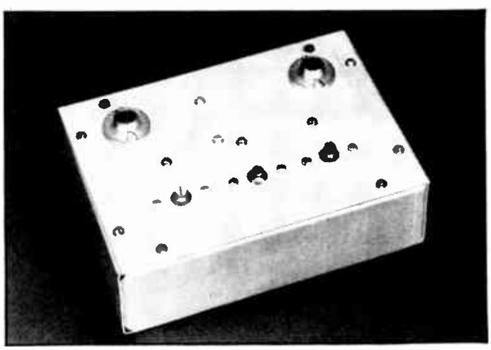


Fig. 10-30 — Television-frequency harmonic filter, for use with coax cable. All parts are mounted on a 5 × 7-inch piece of aluminum, mounted with sheet-metal screws in a 5 by 7 by 2 aluminum chassis which serves as a shield.

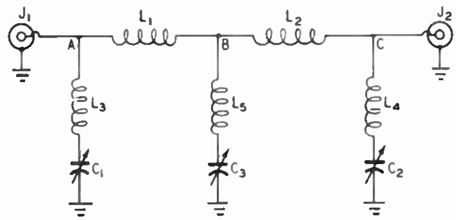


Fig. 10-31 — Circuit diagram of the harmonic filter. It provides two high-attenuation points which may be placed in television channels employed in the locality in which the filter is to be used.

J_1, J_2 — Panel-type coaxial connectors.
C_1, C_2 — 35- or 50-μμfd. variable; see data below. (Millen 22035 or 22050).
C_3 — 100-μμfd. variable (Millen 22100).

Coil and Capacitance Data

For 50-ohm cable, maximum rejection in Channels 2 and 4:
- C_1, C_2 — 12 μμfd.
- C_3 — 106 μμfd. (Condenser specified above has sufficient capacitance.
- L_1, L_2 — 5 turns No. 12, 1/2-inch inside diameter, length 5/8 inch.
- L_3, L_4 — 4 turns No. 12, 1/2-inch inside diameter, length 9/16 inch.
- L_5 — 1 turn No. 12, 1/2-inch inside diameter, length 3/8 inch.

For 50-ohm cable, maximum rejection in Channels 3 and 6:
- C_1, C_2 — 38 μμfd.
- C_3 — 90 μμfd.
- L_1, L_2 — 5 turns No. 12, 1/2-inch inside diameter, length 7/8 inch.
- L_3, L_4 — 4 turns No. 12, 1/2-inch inside diameter, length 11/16 inch.
- L_5 — 1 turn No. 12, 3/8-inch inside diameter, length 1/2 inch.

For 75-ohm cable, maximum rejection in Channels 2 and 4:
- C_1, C_2 — 28 μμfd.
- C_3 — 71 μμfd.
- L_1, L_2 — 7 turns No. 12, 1/2-inch inside diameter, length 3/4 inch.
- L_3, L_4 — 6 turns No. 12, 1/2-inch inside diameter, length 11/16 inch.
- L_5 — 3 turns No. 12, 3/8-inch inside diameter, length 9/16 inch.

For 75-ohm cable, maximum rejection in Channels 3 and 6:
- C_1, C_2 — 25 μμfd.
- C_3 — 66 μμfd.
- L_1, L_2 — 7 turns No. 12, 1/2-inch inside diameter, length 13/16 inch.
- L_3, L_4 — 6 turns No. 12, 1/2-inch inside diameter, length 3/4 inch.
- L_5 — 2 turns No. 12, 3/8-inch inside diameter, length 3/4 inch.

Coil lengths in all cases measured between centers of wire at ends.

cannot be made conveniently, a fairly good adjustment can be obtained by short-circuiting point *A* to the common ground plate (use the shortest possible connection) and adjusting C_1 so that a grid-dip meter coupled to L_3 shows the circuit to be resonant at 57 Mc. for a Channel 2 filter, or at 63 Mc. for a Channel 3 filter. Adjust C_2 similarly with the grid-dip meter coupled to L_4 and point *C* shorted to ground. Then short point *B* to ground at the hole in the shield, Fig. 10-32, couple the grid-dip meter to L_5, and adjust C_3 to 71 Mc. for a Channel 4 filter or to 85 Mc. for a Channel 6 filter. These adjustments usually will provide good average attenuation in the two channels. Should actual interference be caused a more exact adjustment, made while watching the television picture, should result in a considerable increase in attenuation.

The cut-off frequencies in both types of filters are well above 30 Mc., and so the filter should have no effect on the performance of the antenna coupling system at frequencies below 30 Mc. If inserting the filter in the line causes the loading on the final stage to change, it is an indication that the coax line is operating at an s.w.r. greater than 1. Optimum results will be secured when the line is first matched as closely as possible so that it operates at a low s.w.r. A bridge-type indicator such as is described in Chapter Sixteen is excellent for determining the s.w.r. and showing the effect of matching adjustments.

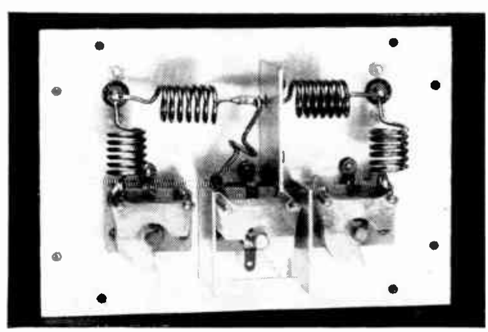


Fig. 10-32 — Construction of the harmonic filter. Dimensions should be followed fairly closely for optimum results. The center-to-center distance between the coax connectors is 1 1/2 inches. Mounting centers of the variable condensers are on a line 2 1/4 inches below and parallel to a line through the centers of the coax fittings.

Antenna-Coupler Construction

The apparatus used to cancel line reactance and match the line resistance to the transmitter is commonly called an "antenna coupler" or "antenna tuner." (The name is really a misnomer, because the coupling and tuning equipment at the input end of the line does not have any effect on the antenna itself; if there is any antenna tuning to be done it must be done at the antenna, independently of the line.) The design principles and the important constructional points have been covered earlier in this chapter; in this section we show a few examples of typical construction.

Bearing in mind the precautions mentioned earlier as to maintaining balance in parallel-conductor transmission lines, it is usually good practice to install the coupling equipment close to the point where the line enters the station. This is a simple matter when the tuning equipment is link-coupled to the trans-

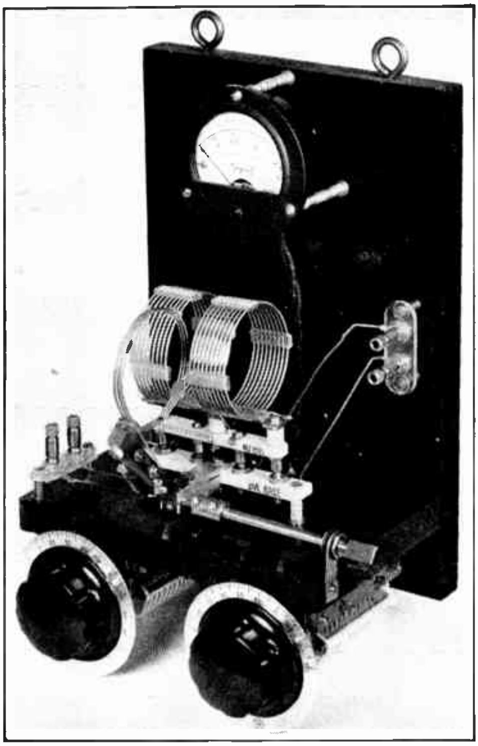


Fig. 10-33 — A wall-mounting antenna coupler for medium-power transmitters. This unit provides a choice of either series or parallel tuning for resonant feeders. Standard transmitting coils of the variable-link type are used.

mitter, since there are no particular restrictions on the length of the link that can be used. However, if the link line is fairly long it should be treated as a transmission line rather than merely as a means of providing mutual inductance between two separated coils. In such a case it is advisable to have variable coupling at both ends of the link. This permits matching the link line to the line tank circuit, and once the match is obtained the power output of the transmitter can be varied by changing the coupling at the transmitter tank. If the link line is not properly matched its current may be excessive, leading to unnecessary power loss.

The most desirable form of link line is coaxial cable. Properly handled, its losses are low; and since it is shielded it can be on or near metal objects with impunity.

SERIES-PARALLEL COUPLER FOR WALL MOUNTING

Fig. 10-33 shows a link-coupled coupler designed for series or parallel tuning of a resonant line. It is suitable for transmitters having a power output in the neighborhood of 250 watts. A higher-power version easily could be made using a similar layout, but substituting heavier coils and condensers with greater plate spacing.

As shown in Fig. 10-34, the change from series to parallel tuning is made by means of jumpers and extra pins on the coil plug bar. A separate coil is used for each band, and after determining which should be used, series or parallel tuning, on a particular band, the jumpers may be installed permanently or left off as required. The tuning condensers specified, together with a set of standard plug-in transmitting coils, should provide adequate coupling if the transmission-line length is such as to bring a voltage or current loop near the input end.

The unit is mounted on an 8 × 12 × ⅞-inch board for hanging on the wall in any convenient location near the entrance point of the feeders. The 2.5-ampere r.f. ammeter is mounted centrally by long wood screws through spacers at the top of the unit. A short length of twisted pair connects it to the thermocouple, secured in a horizontal position at the bottom of the backboard. The tuning condensers are mounted on the underside of a 4-inch shelf extending the width of the unit. Atop the shelf, the jack bar for the coil is supported on pillars by wood screws. An extension shaft to vary the degree of coupling is supported by a bushing fastened to a short strip of brass at the right of the shelf. A short length of 300-ohm ribbon (coaxial cable can be used instead) connects the input terminals to the movable link, while the output terminals are located at the middle right of the backboard. Two screw eyes at the top permit the unit to be hung from screws or nails in the wall.

RACK-MOUNTING SERIES-PARALLEL COUPLER

The rack-mounting coupling unit shown in Fig. 10-35 is suitable for power outputs of 25 to 50 watts, and provides either series or parallel tuning for resonant lines. Separate condensers are used for this purpose, and while

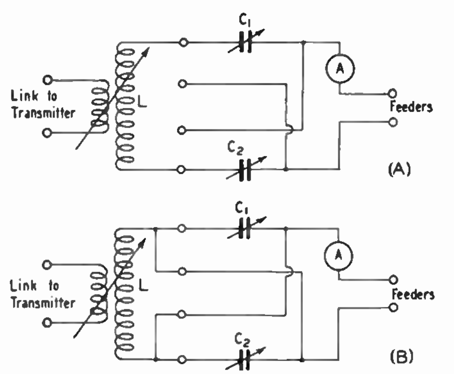


Fig. 10-34 — Circuit diagram of an antenna coupler for use with a medium-power transmitter. A — Series tuning. B — Parallel tuning.

C_1, C_2 — 100-μμfd. single section variable, 0.070-inch spacing (Cardwell MT-100-GS).
L — B & W BVL series.
A — 0–2.5 thermocouple r.f. ammeter.

three are required, this system has the advantage that no switching is necessary when changing from series to parallel tuning. It is also possible to cover a somewhat wider range of line input impedances with parallel tuning because the series condensers can be used to help cancel out inductive reactance that cannot be handled by the parallel circuit alone.

The coupler is mounted on a 5¼ × 19-inch panel. The parallel condenser, C_1, is in the center, with C_2 and C_3 on either side. The variable condensers are mounted on National GS-1 stand-off insulators which are fastened to the condenser tie-rods by means of machine screws with the heads cut off. Small ceramic shaft couplings are used to insulate the control knobs from the condenser shafts.

Clips with flexible leads attached are provided for the parallel condenser, C_1, so that the sections may be used either in series or parallel to form either a high-C or low-C tank circuit, as required. When the high-C tank is necessary the two stators are connected together by means of the clips, as indicated by the dotted lines in the circuit diagram, Fig. 10-36. When the two sections are connected in series for low-C operation the breakdown voltage is increased.

Two sets of variable condensers are suggested in the list of parts. The smaller receiving-type condensers with 0.03-inch air gap are satisfactory for transmitter power outputs up to 50 watts. The larger condensers, with 0.045-inch spacing, are required for transmitter outputs of the order of 100 watts.

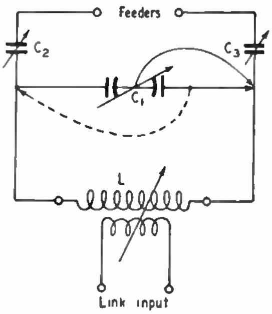


Fig. 10-36 — Circuit of the rack-mounting antenna tuner for use with transmitters having final amplifiers that are operated at less than 1000 volts on the plate.

All coils are 1⅞ inches in diameter and 2¼ inches long, with the variable link located at the center. For series tuning, use the coil specified for the next-higher frequency band, which will be approximately correct.

C_1 — 100 μμfd. per section, 0.045-inch spacing (National TMK-100-D) for high voltages; receiving type for low voltages (Hammarlund MCD-100).

C_2, C_3 — 250 μμfd., 0.026-inch spacing (National TMS-250) for high voltages; receiving type for low voltages (Hammarlund MC-250).

L — B & W JVL-series coils. Approximate dimensions for parallel tuning for each band are as follows:
3.5-Mc. band — 40 turns No. 20.
7-Mc. band — 24 turns No. 16.
14-Mc. band — 14 turns No. 16.
28-Mc. band — 8 turns No. 16.

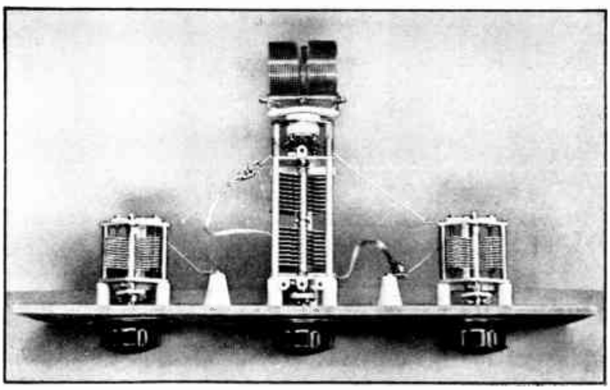
BANDSWITCHING UNIVERSAL COUPLER

The coupling unit shown in Figs. 10-37 and 10-39 is of the "universal" type discussed earlier. It is a bandswitching unit using commercially-available coils. Provision is made for switching either capacitance or inductance across the transmission line to compensate for its input reactance. Impedance matching is achieved by tapping the tank coils at the proper points.

In the circuit diagram, Fig. 10-38, only one set of coils is shown. For other bands the connections shown for L_1 and L_2 would be duplicated. Bandswitching is accomplished by a five-gang switch, S_1. Compensating reactances can be switched in or out of the circuit by S_2. The coupling links, L_2, are the shielded type using coaxial cable described earlier in this chapter (Fig. 10-28).

The coupler is wholly supported by a 7 × 19-inch relay-rack panel. The variable condensers are mounted from the panel by small stand-off insulators, and insulated couplings are used between the condenser shafts and the National Type AM dials. The tank condenser, C_1, is mounted at the right-hand end of the panel with the bandswitch, S_1, to its left. The four coils are grouped around the bandswitch, with the 28-Mc. coil placed so that the leads to it are the shortest. The coils are Millen 44000 series with the plug bases removed from the

Fig. 10-35 — Rack-mounted coupler for low-power transmitters. This unit uses three variable condensers to provide either series or parallel tuning without condenser switching.



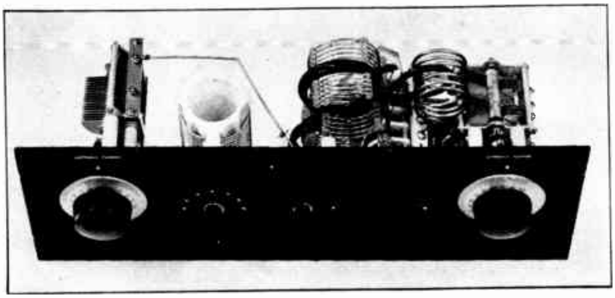


Fig. 10-37 — Bandswitching universal-type coupler for parallel-conductor lines. This unit can be used with transmitters having power outputs of the order of 100 watts.

3.5-, 7- and 14-Mc. coils. It is not practicable to remove the base from the 28-Mc. coil because it does not have the polystyrene supporting strip that is part of the lower-frequency coil

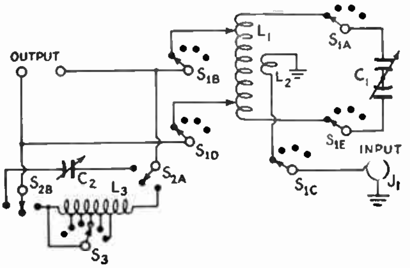


Fig. 10-38 — Circuit diagram of the band-switching coupler. In this diagram the ground symbol indicates points that are connected together. Wiring to coils is shown for one band only, to avoid complicating the diagram: the wiring for other coils is identical.

C_1 — 100-μμfd.-per-section variable (Cardwell MR-100-BD).
C_2 — 335-μμfd. variable (Cardwell MR-335-BS).
L_1 — Millen 44000-series coils (see text).
L_2 — Shielded link; one turn for 28 and 14 Mc.; 2 turns for 7 and 3.5 Mc.
L_3 — 26 turns No. 12 on 2½-inch diameter form (National XR-10A), 7 turns per inch. Tapped 8, 14, 18, 22 and 24 turns from end to which arm of S_3 is connected.
J_1 — Coaxial-cable connector (Amphenol).
S_1 — 5-section 4-position ceramic wafer switch (Centralab 2546).
S_2 — 2-section 4-position ceramic wafer switch (Centralab 2543).
S_3 — 1-section 6-position ceramic wafer switch (Centralab 2501).

assemblies. The coils are partly supported by the wiring to the switch and partly by the polystyrene plate mounted on the back of the switch. The ends of the coil mounting strips are cemented into holes cut in the plate.

The compensating condenser, C_2, is mounted at the left-hand end of the panel. L_3 is mounted vertically to its right, with S_3 directly in front of it on the panel. S_2 is mounted centrally on the panel. The output terminals to the line are mounted above S_3. The link input terminal is a coaxial cable socket mounted on a small bracket in the lower right-hand corner.

The link coils, L_2, are supported by the wiring, and the coupling is changed by bending the link into or out of its associated tank coil. Since the links fit rather tightly in the tank coils, the pressure helps hold them in place once the proper coupling is determined. The link shields are all connected together and to the input connector; the inner conductors go to the switch contacts. The link coils are made from RG-59/U cable.

With the coils and condensers specified, this coupler can handle power outputs of the order of 100 to 150 watts. The method of adjustment is covered earlier in this chapter.

A WIDE-RANGE ANTENNA COUPLER

The photograph of Fig. 10-40 shows the constructional details of a wide-range antenna coupler suitable for use with high-power transmitters. Various combinations of parallel and series tuning, with high- and low-C tanks and high- and low-impedance outputs, are available. Diagrams of the various circuit combinations possible with this arrangement are given in Fig. 10-41.

A separate coil is used for each band, and the desired connections for series or parallel tuning with high or low C, or for low-impedance output

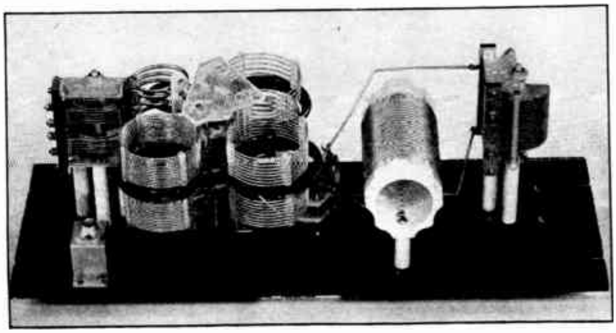


Fig. 10-39 — Rear view of the band-switching coupler. Details of coil mountings are shown in this view.

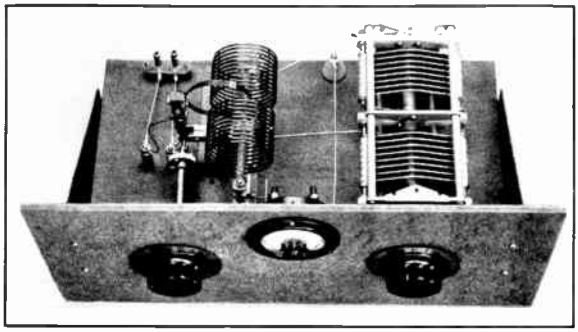


Fig. 10-40 — Wide-range antenna coupler. The unit is assembled on a metal chassis measuring 10 × 17 × 2 inches, with a panel $8\frac{3}{4}$ × 19 inches in size. The variable condenser is a split-stator unit with a capacitance of 200 μμfd. per section and 0.07-inch plate spacing (Johnson 200ED30). The plug-in coils are the B & W TVL series. The r.f. ammeter has a 1-ampere scale.

with high or low C, are automatically made when the coil is plugged in. Coil connections to the pins for various circuit arrangements are shown in Fig. 10-41.

The tuning condenser specified, together with a set of standard plug-in transmitting coils, should cover nearly all coupling conditions likely to be encountered.

Because the switching connections require the use of a central pin, a slight alteration in the B & W coil-mounting unit is required. The central link-mounting unit should be removed from the jack-bar and an extra jack placed in the central hole thus made available. The link assembly should then be mounted on a 2-inch cone insulator to one side of the jack bar.

Correspondingly, the central nut on each coil plug base must be removed and a Johnson tapped plug, similar to those furnished with the coils, substituted. An extension shaft may then be fitted on the link shaft and a control brought out to a knob on the panel.

The split-stator tank condenser is mounted by means of angle brackets on four 1-inch cone-type ceramic insulators, and an insulated flexible coupling is provided for the shaft.

If desired, the coils may be wound with fixed links on ceramic transmitting coil forms. The links should be provided with flexible leads which can be plugged into a pair of jack-top insulators mounted near the coil jack strip, unless a special mounting is made providing for seven connections.

The unit as described should be satisfactory for transmitters having an output of 500 watts with plate modulation and somewhat more on c.w. For higher-power 'phone, a tank condenser with larger plate spacing should be used.

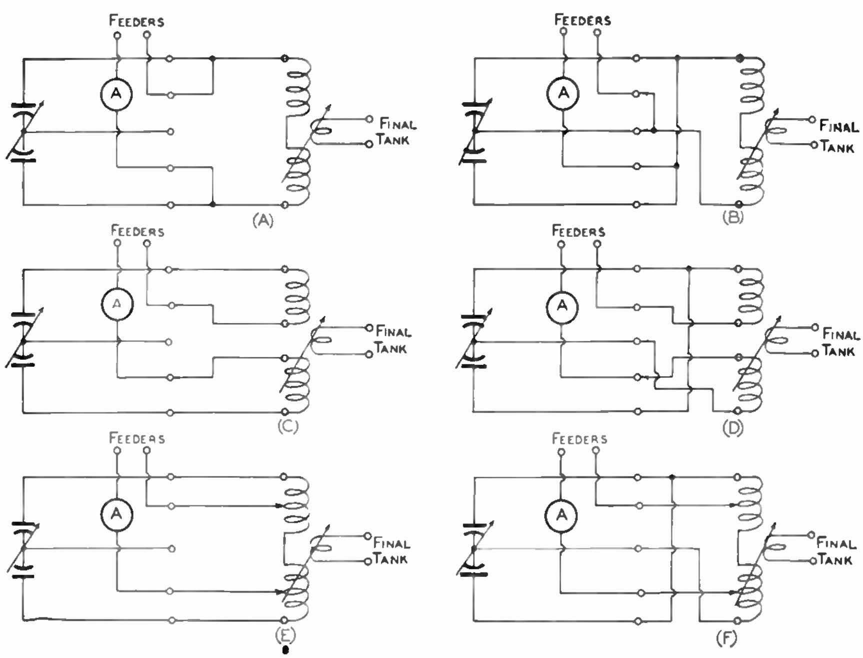


Fig. 10-41 — Circuit diagram of the wide-range rack-type antenna coupler. A — Parallel tuning, low C. B — Parallel tuning, high C. C — Series tuning, low C. D — Series tuning, high C. E — Parallel tank, low-impedance output, low C. F — Parallel tank, low-impedance output, high C. After the inductance required for each of the various bands has been determined experimentally, the connections to the coils can be made permanent. Then it will be necessary only to plug in the right coil for each band, tune the condenser for resonance, and adjust the link loading.

Antennas

In selecting the type of antenna to use, the propagation characteristics of the frequency band or bands to be used should be given due consideration. These are outlined in Chapter Four. In general, antenna construction and location become more critical and important on the higher frequencies. To some extent on 14 Mc. and to an even greater degree on the higher bands, the angle of radiation should be as low as possible for good results over long-distance paths. On any one band, however, an antenna well-suited for long-distance work is not likely to be as suitable for short-haul contacts as some other type of antenna. Important properties of an antenna or antenna system are its polarization, angle of radiation, impedance, directivity and gain.

Polarization

The polarization of a straight-wire antenna is its position with respect to the earth. That is, a vertical wire transmits vertically-polarized waves and a horizontal antenna generates horizontally-polarized waves in its direction of maximum radiation (broadside). The wave from an antenna in a slanting position contains both horizontal and vertical components.

Angle of Radiation

The **wave angle** (or vertical angle) at which an antenna radiates best is determined by its polarization, height above ground, and the nature of the ground. Radiation is not all at one well-defined angle, but rather is generally dispersed over a more or less large angular region, depending upon the type of antenna. The angle is measured in a vertical plane with respect to a tangent to the earth at that point.

Impedance

The impedance of the antenna at any point is the ratio of the voltage to the current at that point. It is important in connection with feeding power to the antenna, since it constitutes the load represented by the antenna. It is a pure resistance only at current loops (maxima) and nodes (minima) on resonant antennas. The antenna impedance is high at the current node and low at the current loop.

Directivity

All antennas radiate more power in certain directions than in others. This characteristic, called *directivity*, must be considered in three dimensions, since directivity exists in the vertical plane as well as in the horizontal plane. Thus the directivity of the antenna will affect the wave angle as well as the actual compass directions in which maximum transmission takes place.

Current

The **field strength** produced by an antenna is proportional to the current flowing in it. When there are standing waves on an antenna, the parts of the wire carrying the higher current have the greater radiating effect. All resonant antennas have standing waves — only terminated types, like the terminated rhombic and terminated "V," have substantially uniform current along their lengths.

Power Gain

The ratio of power required to produce a given field strength, with a "comparison" antenna, to the power required to produce the same field strength with a specified type of antenna is called the **power gain** of the latter antenna. The field is measured in the optimum direction of the antenna under test. In amateur work, the comparison antenna is generally a half-wave antenna at the same height and having the same polarization as the antenna under consideration. Power gain usually is expressed in decibels.

Front-to-Back Ratio

In unidirectional beams (antenna systems with maximum radiation in only one direction) the front-to-back ratio is the ratio of power radiated in the maximum direction to power radiated in the opposite direction. It is also a measure of the reduction in received signal when the beam direction is changed from that for maximum response to the opposite direction. Front-to-back ratio is usually expressed in decibels.

Ground Effects

The radiation pattern of any antenna that is many wavelengths distant from the ground and all other objects is called the **free-space pattern** of that antenna. The free-space pattern of an antenna is almost impossible to obtain in practice, except in the v.h.f. and u.h.f. ranges. Below 30 Mc., the location of the antenna with respect to ground plays an important part in determining the actual radiation pattern of the antenna.

When any antenna is near the ground the free-space pattern is modified by reflection of radiated waves from the ground, so that the actual pattern is the resultant of the free-space pattern and ground reflections. This resultant is dependent upon the height of the antenna, its position or orientation with respect to the surface of the ground, and the electrical characteristics of the ground. The effect of a perfectly-reflecting ground is such that the

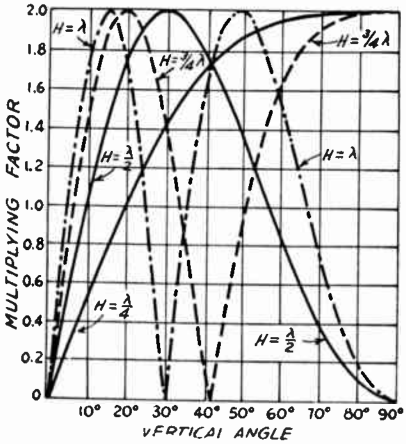


Fig. 10-42 — Effect of ground on radiation of horizontal antennas at vertical angles for four antenna heights. This chart is based on perfectly-conducting ground.

original free-space field strength may be multiplied by a factor which has a maximum value of 2, for complete reinforcement, and having all intermediate values to zero, for complete cancellation. These reflections only affect the radiation pattern in the vertical plane — that is, in directions upward from the earth's surface — and not in the horizontal plane, or the usual geographical directions.

Fig. 10-42 shows how the multiplying factor varies with the vertical angle for several representative heights for horizontal antennas. As the height is increased the angle at which complete reinforcement takes place is lowered, until for a height equal to one wavelength it occurs at a vertical angle of 15 degrees. At still greater heights, not shown on the chart, the first maximum will occur at still smaller angles.

Radiation Angle

The vertical angle, or angle of radiation, is of primary importance, especially at the higher frequencies. It is advantageous, therefore, to erect the antenna at a height that will take advantage of ground reflection in such a way as to reinforce the space radiation at the most desirable angle. Since low radiation angles usually are desirable, this generally means that the antenna should be high — at least one-half wavelength at 14 Mc., and preferably three-quarters or one wavelength; at least one wavelength, and preferably higher, at 28 Mc. and the very-high frequencies. The physical height required for a given height in wavelengths decreases as the frequency is increased, so that good heights are not impracticable; a half-wavelength at 14 Mc. is only 35 feet, approximately, while the same height represents a full wavelength at 28 Mc. At 7 Mc. and lower frequencies the higher radiation angles are effective, so that again a reasonable antenna height is not difficult of attainment. Heights between 35 and 70 feet are suitable for all bands, the higher figures being preferable.

Imperfect Ground

Fig. 10-42 is based on ground having perfect conductivity, whereas the actual earth is not a perfect conductor. The principal effect of actual ground is to make the curves inaccurate at the lowest angles; appreciable high-frequency radiation at angles smaller than a few degrees is practically impossible to obtain over horizontal ground. Above 15 degrees, however, the curves are accurate enough for all practical purposes, and may be taken as indicative of the sort of result to be expected at angles between 5 and 15 degrees.

The effective ground plane — that is, the plane from which ground reflections can be considered to take place — seldom is the actual surface of the ground but is a few feet below it, depending upon the character of the soil.

Impedance

Waves that are reflected directly upward from the ground induce a current in the antenna in passing, and, depending on the antenna height, the phase relationship of this induced current to the original current may be such as either to increase or decrease the total current in the antenna. For the same power input to the antenna, an increase in current is equivalent to a decrease in impedance, and vice versa. Hence, the impedance of the antenna varies with height. The theoretical curve of variation of radiation resistance for an antenna above perfectly-reflecting ground is shown in Fig. 10-43. The impedance approaches the free-space value as the height becomes large, but at low heights may differ considerably from it.

Choice of Polarization

Polarization of the transmitting antenna is generally unimportant on frequencies between 3.5 and 30 Mc. However, the question of whether the antenna should be installed in a horizontal or vertical position deserves consideration for other reasons. A vertical half-wave or quarter-wave antenna will radiate

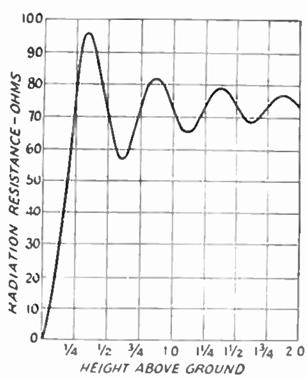


Fig. 10-43 — Theoretical curve of variation of radiation resistance for a half-wave horizontal antenna, as a function of height in wavelength above perfectly-reflecting ground.

equally well in all *horizontal* directions, so that it is substantially nondirectional, in the usual sense of the word. If installed horizontally, however, the antenna will tend to show directional effects, and will radiate best in the direction at right angles, or broadside, to the wire. The radiation in such a case will be least in the direction toward which the wire points.

The vertical angle of radiation also will be affected by the position of the antenna. If it were not for ground losses at high frequencies, the vertical half-wave antenna would be preferred because it would concentrate the radiation horizontally.

The Half-Wave Antenna

The fundamental form of antenna is a single wire whose length is approximately equal to half the transmitting wavelength. It is the unit from which many more-complex forms of antennas are constructed. It is variously known as a **half-wave dipole, half-wave doublet,** or **Hertz** antenna.

The length of a half-wavelength in space is:

$$\textit{Length (feet)} = \frac{492}{\textit{Freq. (Mc.)}} \qquad \textbf{(10-H)}$$

The actual length of a half-wave antenna will not be exactly equal to the half-wave in space, but depends upon the thickness of the conductor in relation to the wavelength as shown in Fig. 10-44, where K is a factor that must be multiplied by the half-wavelength in free space to obtain the resonant antenna length. An additional shortening effect occurs with wire antennas supported by insulators at the ends because of the capacitance added to the system by the insulators **(end effect).** The following formula is sufficiently accurate for wire antennas at frequencies up to 30 Mc.:

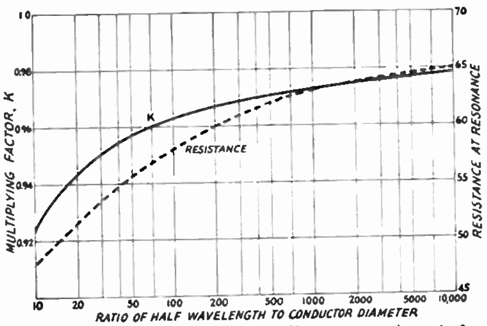


Fig. 10-44 — Effect of antenna diameter on length for half-wave resonance, shown as a multiplying factor, K, to be applied to the free-space half-wavelength (Equation 10-H). The effect of conductor diameter on the impedance measured at the center also is shown.

$$\textit{Length of half-wave antenna (feet)} = \frac{492 \times 0.95}{\textit{Freq. (Mc.)}} = \frac{468}{\textit{Freq. (Mc.)}} \qquad \textbf{(10-I)}$$

Example: A half-wave antenna for 7150 kc. (7.15 Mc.) is $\frac{468}{7.15} = 65.45$ feet, or 65 feet 5 inches.

Above 30 Mc. the following formulas should be used, particularly for antennas constructed from rod or tubing. K is taken from Fig. 10-44.

$$\textit{Length of half-wave antenna (feet)} = \frac{492 \times K}{\textit{Freq. (Mc.)}} \qquad \textbf{(10-J)}$$

$$\textit{or length (inches)} = \frac{5905 \times K}{\textit{Freq. (Mc.)}} \qquad \textbf{(10-K)}$$

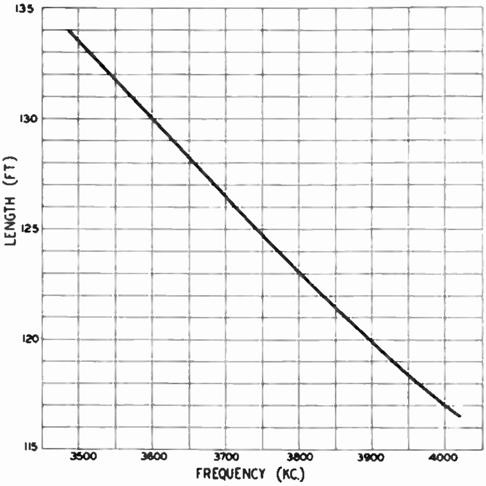
Example: Find the length of a half-wavelength antenna at 29 Mc., if the antenna is made of 2-inch diameter tubing. At 29 Mc., a half-wavelength in space is $\frac{492}{29} = 16.97$ feet, from Eq. 10-H. Ratio of half-wavelength to conductor diameter (changing wavelength to inches) is $\frac{16.97 \times 12}{2} = 101.8$. From Fig. 10-44, $K = 0.963$ for this ratio. The length of the antenna, from Eq. 10-J, is $\frac{492 \times 0.963}{29} = 16.34$ feet, or 16 feet 4 inches. The answer is obtained directly in inches by substitution in Eq. 10-K: $\frac{5905 \times 0.963}{29} = 196$ inches.

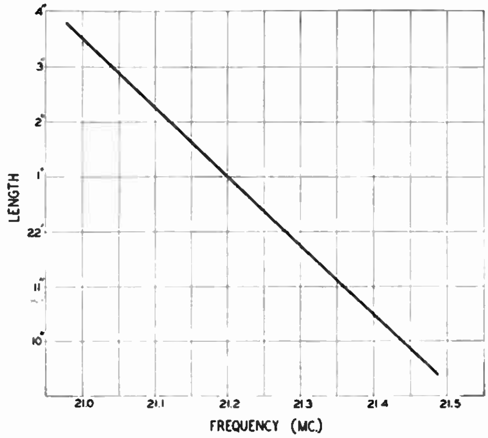
Current and Voltage Distribution

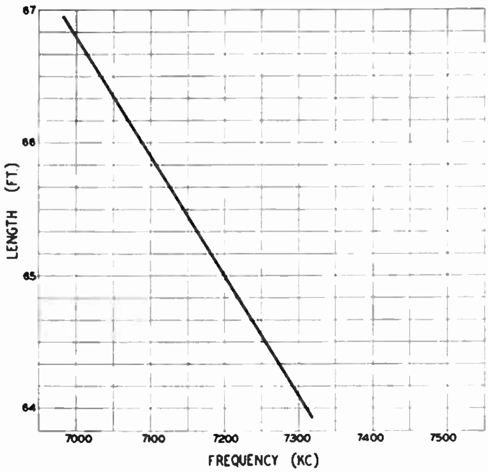
When power is fed to such an antenna, the current and voltage vary along its length. The current is maximum at the center and nearly zero at the ends, while the opposite is true of the r.f. voltage. The current does not actually reach zero at the current nodes, because of the end effect; similarly, the voltage is not zero at its node because of the resistance of the antenna, which consists of both the r.f. resistance of the wire (*ohmic resistance*) and the **radiation resistance.** The radiation resistance is an *equivalent* resistance, a convenient conception to indicate the radiation properties of an antenna. The radiation resistance is the equivalent resistance that would dissipate the power the antenna radiates, with a current flowing in it equal to the antenna current at a current loop (maximum). The ohmic resistance of a half-wavelength antenna is ordinarily small enough, in comparison with the radiation resistance, to be neglected for all practical purposes.

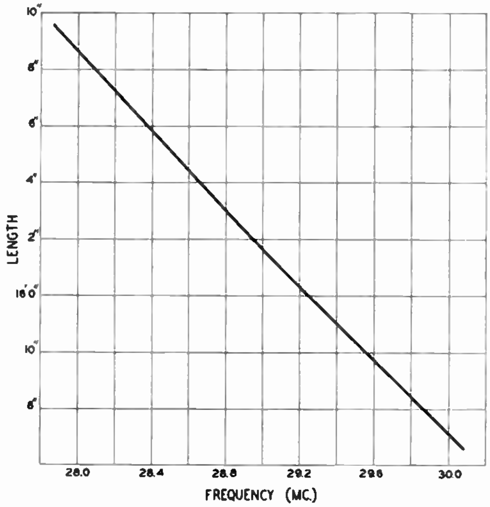
Impedance

The radiation resistance of an infinitely-thin half-wave antenna in free space — that is, sufficiently removed from surrounding objects so that they do not affect the antenna's characteristics — is 73 ohms, approximately. The value under practical conditions is commonly taken to be in the neighborhood of 70 ohms. It is pure resistance, and is measured at the center of the antenna. The impedance









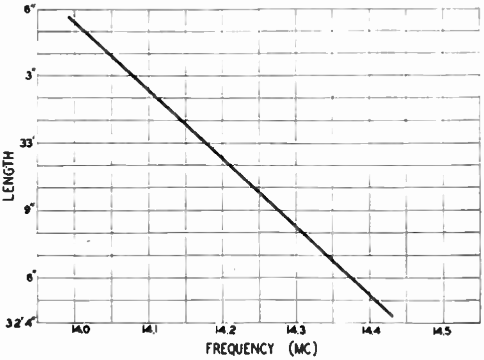


Fig. 10-45 — The above charts, based on Eq. 10-I, can be used to determine the length of a half-wave antenna of wire.

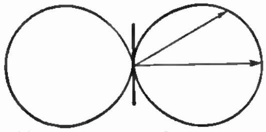


Fig. 10-46 — The free-space radiation pattern of a half-wave antenna. The antenna is shown in the vertical position. This is a cross-section of the solid pattern described by the figure when rotated on its vertical axis. The "doughnut" form of the solid pattern can be more easily visualized by imagining the drawing glued to a piece of cardboard, with a short length of wire fastened on it to represent the antenna. Twirling the wire will give a visual representation of the solid radiation pattern.

is minimum at the center, where it is equal to the radiation resistance, and increases toward the ends. The actual value at the ends will depend on a number of factors, such as the height, the physical construction, the insulators at the ends, and the position with respect to ground.

Conductor Size

The impedance of the antenna also depends upon the diameter of the conductor in relation to the wavelength, as shown in Fig. 10-44. If the diameter of the conductor is made large, the capacitance per unit length increases and the inductance per unit length decreases. Since the radiation resistance is affected relatively little, the decreased L/C ratio causes the Q of the antenna to decrease, so that the resonance curve becomes less sharp. Hence, the antenna is capable of working over a wide frequency range. This effect is greater as the diameter is increased, and is a property of some importance at the very-high frequencies where the wavelength is small.

Radiation Characteristics

The radiation from a half-wave antenna is not uniform in all directions but varies with the angle with respect to the axis of the wire. It is most intense in directions perpendicular to the wire and zero along the direction of the wire, with intermediate values at intermediate angles. This is shown by the sketch of Fig. 10-46, which represents the radiation pattern in free space. The relative intensity of radiation is proportional to the length of a line drawn from the center of the figure to the perimeter. If the antenna is vertical, as shown in the figure, then the field strength will be uniform in all horizontal directions; if the antenna is horizontal, the relative field strength will depend upon the direction of the receiving point with respect to the direction of the antenna wire. The variation in radiation at various vertical angles from a half-wavelength horizontal antenna is indicated in Figs. 10-47 and 10-48.

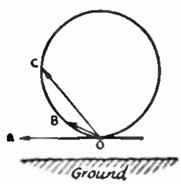


Fig. 10-47 — Illustrating the importance of vertical angle of radiation in determining antenna directional effects. Ground reflection is neglected in this drawing of the free-space field pattern of a horizontal antenna.

FEEDING THE HALF-WAVE ANTENNA

Direct Feed

If possible, it is advisable to locate the antenna at least a half-wavelength from the transmitter and use a transmission line to carry the power from the transmitter to the antenna. However, in many cases this is impossible, particularly on the lower frequencies, and direct feed must be used. Three examples of direct feed are shown in Fig. 10-49. In the method shown at A, C_1 and C_2 should be about 150 $\mu\mu$fd. each for the 3.5-Mc. band, 75 $\mu\mu$fd. each at 7 Mc., and proportionately smaller at the higher frequencies. The antenna coil

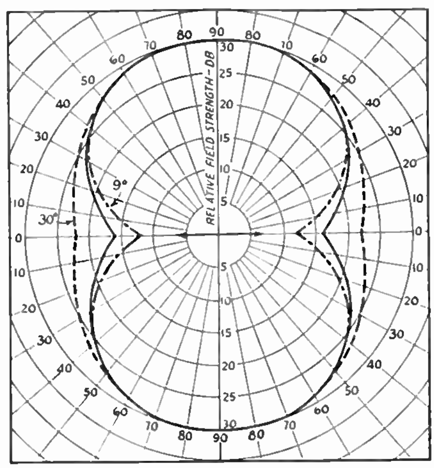


Fig. 10-48 — Horizontal pattern of a horizontal half-wave antenna at three vertical radiation angles. The solid line is relative radiation at 15 degrees. Dotted lines show deviation from the 15-degree pattern for angles of 0 and 30 degrees. The patterns are useful for shape only, since the amplitude will depend upon the height of the antenna above ground and the vertical angle considered. The patterns for all three angles have been proportioned to the same scale, but this does not mean that the maximum amplitudes necessarily will be the same. The arrow indicates the direction of the horizontal antenna wire.

connected between them should resonate to 3.5 Mc. with about 60 or 70 $\mu\mu$fd., for the 80-meter band, for 40 meters it should resonate with 30 or 35 $\mu\mu$fd., and so on. The circuit is adjusted by using loose coupling between the antenna coil and the transmitter tank coil and adjusting C_1 and C_2 until resonance is indicated by an increase in plate current. The coupling between the coils should then be increased until proper plate current is drawn. It may be necessary to reresonate the transmitter tank circuit as the coupling is increased, but the change should be small.

The circuits in Fig. 10-49B and C are used when only one end of the antenna is accessible. In B, the coupling is adjusted by moving the

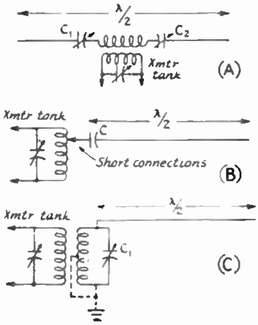


Fig. 10-49 — Methods of directly exciting the half-wave antenna. A, current feed, series tuning; B, voltage feed, capacitive coupling; C, voltage feed, with inductively-coupled antenna tank. In A, the coupling circuit is not included in the effective electrical length of the antenna-system proper.

tap toward the "hot" or plate end of the tank coil — the condenser C may be of any convenient value that will stand the voltage, and it doesn't have to be variable. In the circuit at C, the antenna tuned circuit (C_1 and the antenna coil) should be similar to the transmitter tank circuit. The antenna tuned circuit is adjusted to resonance with the antenna connected but with loose coupling to the transmitter. Heavier loading of the tube is then obtained by tightening the coupling between the antenna coil and the transmitter tank coil.

Of the three systems, that at A is preferable because it is a symmetrical system and generally results in less r.f. power "floating" around the shack. The system of B is undesirable because it provides practically no protection against the radiation of harmonics, and it should only be used in emergencies.

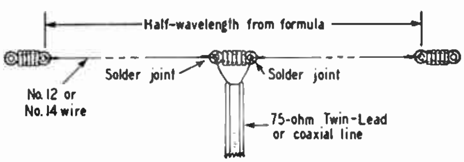


Fig. 10-50 — Construction of a half-wave doublet fed with 75-ohm line. The length of the antenna is calculated from Equation 10-I or Fig. 10-45.

Transmission-Line Feed for Half-Wave Antennas

Since the impedance at the center of a half-wavelength antenna is in the vicinity of 75 ohms, it offers a good match for 75-ohm two-wire transmission lines. Several types are available on the market, with different power-handling capabilities. They can be connected in the center of the antenna, across a small strain insulator to provide a convenient connection point. Coaxial line of 75 ohms impedance can also be used, but it is heavier and thus not as convenient. In either case, the transmission line should be run away at right angles to the antenna for at least one-quarter wavelength, if possible, to avoid current unbalance in the line caused by pick-up from the antenna. The antenna length is calculated from Equation 10-I, for a half-wavelength antenna. When No. 12 or No. 14 enameled wire is used for the antenna, as is generally the case, the length of the wire is the over-all length measured from the loop through the insulator at each end. This is illustrated in Fig. 10-50.

The use of 75-ohm line results in a "flat" line over most of any amateur band. However, by making the half-wave antenna in a special manner, called the **two-wire** or **folded doublet**, a good match is offered for a 300-ohm line. Such an antenna is shown in Fig. 10-51, with another version in Fig. 10-84B. The two differ only in the construction of the antenna

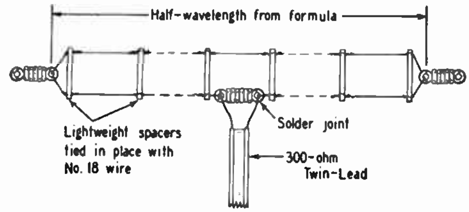


Fig. 10-51 — The construction of an open-wire folded doublet fed with 300-ohm line. The length of the antenna is calculated from Equation 10-I or Fig. 10-45.

proper. The open-wire line shown in Fig. 10-51 is made of No. 12 or No. 14 enameled wire, separated by lightweight spacers of Lucite or other material (it doesn't have to be a *low-loss* insulating material), and the spacing can be on the order of from 4 to 8 inches, depending upon what is convenient and what the operating frequency is. At 14 Mc., 4-inch separation is satisfactory, and 8-inch or even greater spacing can be used at 3.5 Mc.

If a half-wavelength antenna is fed at the center with other than 75-ohm line, or if a folded doublet is fed with other than 300-ohm line, standing waves will appear on the line and coupling to the transmitter may become awkward for some line lengths, as described earlier in this chapter. However, in many cases it is not convenient to feed the half-wave antenna with the correct line (as is the case where multiband operation of the same antenna is desired), and sometimes it is not convenient to feed the antenna at the center. Where multiband operation is desired (to be discussed later) or when the antenna must be

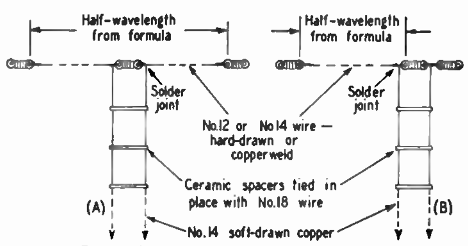


Fig. 10-52 — The antenna can be fed at the center or at the end with an open-wire line. The antenna length is obtained from Equation 10-I or Fig. 10-45.

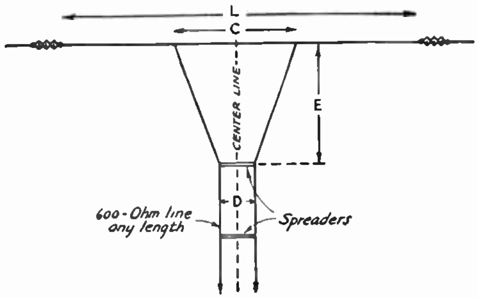


Fig. 10-53 — Delta-matched antenna system. The dimensions C, D, and E are found by formulas given in the text. It is important that the matching section, E, come straight away from the antenna without any bends.

fed at one end by a transmission line, an open-wire line of from 450 to 600 ohms impedance is generally used. The impedance at the end of a half-wavelength antenna is in the vicinity of several thousand ohms, and hence a standing-wave ratio of 4 or 5 is not unusual when the line is connected to the end of the antenna. It is advisable, therefore, to keep the losses in the line as low as possible. This requires the use of ceramic or Micalex feeder spacers, if any appreciable power is used. For low-power installations in dry climates, dry wood spacers that have been boiled in paraffin are satisfactory. Mechanical details of half-wavelength antennas fed with open-wire lines are given in Fig. 10-52. If the power level is low, below 100 watts or so, 300-ohm Twin-Lead can be used in place of the open line.

One method for offering a match to a 600-ohm open-wire line with a half-wavelength antenna is shown in Fig. 10-53. The system is called a **delta match.** The line is "fanned" as it approaches the antenna, to have a gradually-increasing impedance that equals the antenna impedance at the point of connection. The dimensions are fairly critical, but careful measurement before installing the antenna and matching section is generally all that is necessary. The length of the antenna, L, is calculated from Equation 10-I or Fig. 10-45. The length of section C is computed from:

$$C \text{ (feet)} = \frac{118}{\text{Freq. (Mc.)}} \qquad \text{(10-L)}$$

The feeder clearance, E, is found from

$$E \text{ (feet)} = \frac{148}{\text{Freq. (Mc.)}} \qquad \text{(10-M)}$$

Example: For a frequency of 7.1 Mc., the length

$L = \frac{468}{7.1} = 65.91$ feet, or 65 feet 11 inches.

$C = \frac{118}{7.1} = 16.62$ feet, or 16 feet 7 inches.

$E = \frac{148}{7.1} = 20.84$ feet, or 20 feet 10 inches.

Since the equations hold only for 600-ohm line, it is important that the line be close to this value. This requires $4\frac{3}{4}$-inch spaced No. 14 wire, 6-inch spaced No. 12 wire, or $3\frac{3}{4}$-inch spaced No. 16 wire.

Long-Wire Antennas

An antenna will be resonant so long as an integral number of standing waves of current and voltage can exist along its length; in other words, so long as its length is some integral multiple of a half-wavelength. When the antenna is more than a half-wave long it usually is called a long-wire antenna, or a harmonic antenna.

Current and Voltage Distribution

Fig. 10-54 shows the current and voltage distribution along a wire operating at its fundamental frequency (where its length is equal to a half-wavelength) and at its second, third and fourth harmonics. For example, if the fundamental frequency of the antenna is 7 Mc., the current and voltage distribution will be as shown at A. The same antenna excited at 14 Mc. would have current and voltage distribution as shown at B. At 21 Mc., the third harmonic of 7 Mc., the current and voltage distribution would be as in C; and at 28 Mc., the fourth harmonic, as in D. The number of the harmonic is the number of half-waves contained in the antenna at the particular operating frequency.

The polarity of current or voltage in each standing wave is opposite to that in the ad-

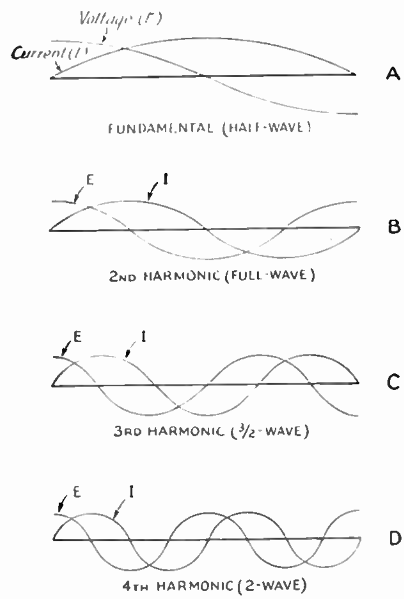


Fig. 10-54 — Standing-wave current and voltage distribution along an antenna when it is operated at various harmonics of its fundamental resonant frequency.

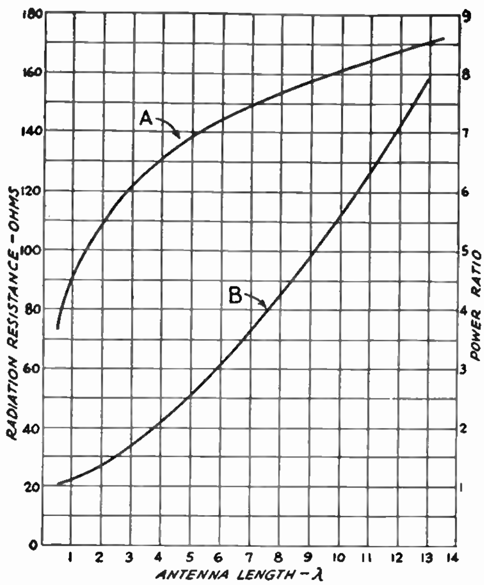


Fig. 10-55 — Curve *A* shows variation in radiation resistance with antenna length. Curve *B* shows power in lobes of maximum radiation for long-wire antennas as a ratio to the maximum radiation for a half-wave antenna.

jacent standing waves. This is shown in the figure by drawing the current and voltage curves successively above and below the antenna (taken as a zero reference line), to indicate that the polarity reverses when the current or voltage goes through zero. Currents flowing in the same direction are *in phase;* in opposite directions, *out of phase.*

It is evident that one antenna may be used for harmonically-related frequencies, such as the various amateur bands. The long-wire or harmonic antenna is the basis of multiband operation with one antenna.

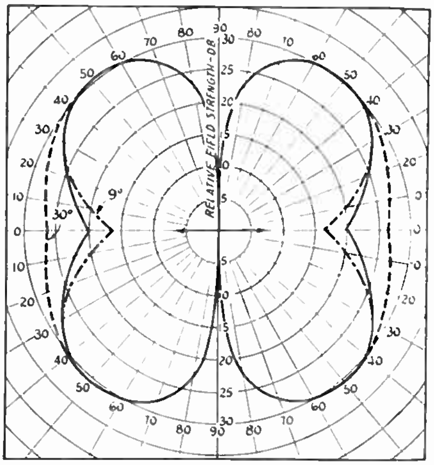


Fig. 10-56 — Horizontal patterns of radiation from a *full-wave* antenna. The solid line shows the pattern for a vertical angle of 15 degrees; dotted lines show deviation from the 15-degree pattern at 9 and 30 degrees. All three patterns are drawn to the same relative scale: actual amplitudes will depend upon the height of the antenna.

Physical Lengths

The length of a long-wire antenna is not an exact multiple of that of a half-wave antenna because the end effects operate only on the end sections of the antenna; in other parts of the wire these effects are absent, and the wire length is approximately that of an equivalent portion of the wave in space. The formula for the length of a long-wire antenna, therefore, is

$$\textit{Length}\ (\text{feet}) = \frac{492\ (N-0.05)}{\textit{Freq.}\ (\text{Mc.})} \qquad \textbf{(10-N)}$$

where N is the number of *half*-waves on the antenna.

Example: An antenna 4 half-waves long at 14.2 Mc. would be $\frac{492\ (4-0.05)}{14.2} = \frac{492 \times 3.95}{14.2}$ = 136.7 feet, or 136 feet 8 inches.

It is apparent that an antenna cut as a half-wave for a given frequency will be slightly off

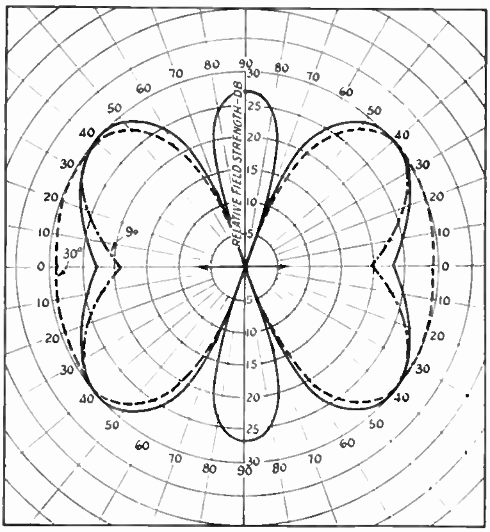


Fig. 10-57 — Horizontal patterns of radiation from an antenna *three half-waves* long. The solid line shows the pattern for a vertical angle of 15 degrees; dotted lines show deviation from the 15-degree pattern at 9 and 30 degrees. Minor lobes coincide for all three angles.

resonance at exactly twice that frequency (the second harmonic), because of the decreased influence of the end effects when the antenna is more than one-half wavelength long. The effect is not very important, except for a possible unbalance in the feeder system and consequent radiation from the feedline. If the antenna is fed in the exact center, no unbalance will occur at any frequency, but end-fed systems will show an unbalance in all but one frequency band, the band for which the antenna is cut.

Impedance and Power Gain

The radiation resistance as measured at a current loop becomes larger as the antenna length is increased. Also, a long-wire antenna radiates more power in its most favorable direction than does a half-wave antenna in its most favorable direction. This power gain is secured at the expense of radiation in other directions. Fig. 10-55 shows how the radiation resistance and the power in the lobe of maximum radiation vary with the antenna length.

Directional Characteristics

As the wire is made longer in terms of the number of half-wavelengths, the directional effects change. Instead of the "doughnut" pattern of the half-wave antenna, the directional characteristic splits up into "lobes" which make various angles with the wire. In general, as the length of the wire is increased the direction in which maximum radiation occurs tends to approach the line of the antenna itself.

Directional characteristics for antennas one wavelength, three half-wavelengths, and two wavelengths long are given in Figs. 10-56, 10-57 and 10-58, for three vertical angles of radiation. Note that, as the wire length increases, the radiation along the line of the antenna becomes more pronounced. Still longer antennas can be considered to have practically "end-on" directional characteristics, even at the lower radiation angles.

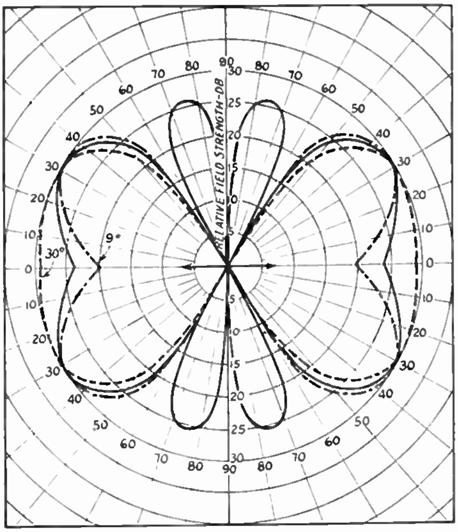


Fig. 10-58 — Horizontal patterns of radiation from an antenna *two wavelengths* long. The solid line shows the pattern for a vertical angle of 15 degrees; dotted lines show deviation from the 15-degree pattern at 9 and 30 degrees. The minor lobes coincide for all three angles.

Methods of Feeding

In a long-wire antenna, the currents in adjacent half-wave sections must be out of phase, as shown in Fig. 10-54. The feeder system must not upset this phase relationship. This requirement is met by feeding the antenna at either end or at any current *loop*. A two-wire feeder cannot be inserted at a current *node*, however, because this invariably brings the currents in two adjacent half-wave sections in phase; if the phase in one section could be reversed, then the currents in the feeders necessarily would have to be in phase and the feeder radiation would not be canceled out.

No point on a long-wire antenna offers a reasonable impedance for a direct match to any of the common types of transmission lines. The most common practice is to feed the antenna at one end or at a current loop with a low-loss open-wire line and accept the resulting standing-wave ratio of 4 or 5. When a better match is required, "stubs" are generally used (described later in this chapter).

Multiband Antennas

As suggested in the preceding section, the same antenna may be used for several bands by operating it on harmonics. When this is done it is necessary to use resonant feeders, since the impedance matching for nonresonant feeder operation can be accomplished only at one frequency unless means are provided for changing the length of a matching section and shifting the point at which the feeder is attached to it.

Furthermore, the current loops shift to a new position on the antenna when it is operated on harmonics, further complicating the feed situation. It is for this reason that a half-wave antenna that is center-fed by a rubber-insulated line is practically useless for harmonic operation; on all even harmonics there is a voltage maximum occurring right at the feed point, and the resultant impedance mismatch is so bad that there is a large standing-wave ratio and consequently high losses arise in the rubber dielectric. It is also wise not to attempt to use a half-wave antenna center-fed with coaxial cable on its harmonics. Higher-impedance solid-dielectric lines such as 300-ohm Twin-Lead may be used, however, provided the power does not exceed a few hundred watts.

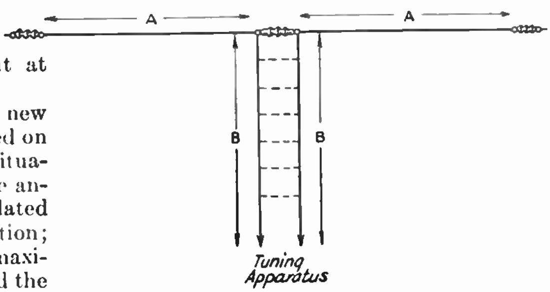


Fig. 10-59 — Practical arrangement of a shortened antenna. The total length, $A + B + B + A$, should be a half-wavelength for the lowest-frequency band, usually 3.5 Mc. See Table 10-III for lengths and tuning data.

TABLE 10-II
Multiband Resonant-Line Fed Antennas

Antenna Length (ft.)	*Feeder Length (ft.)*	*Band*	*Type of Tuning*
With end feed: 120	60	4-Mc. 'phone	series
136	67	3.5-Mc. c.w. 7 Mc. 14 Mc. 28 Mc.	series parallel parallel parallel
134	67	3.5-Mc. c.w. 7 Mc.	series parallel
67	33	7 Mc. 14 Mc. 28 Mc.	series parallel parallel
With center feed: 137	67	3.5 Mc. 7 Mc. 14 Mc. 28 Mc.	parallel parallel parallel parallel
67.5	34	7 Mc. 14 Mc. 28 Mc.	parallel parallel parallel

The antenna lengths given represent compromises for harmonic operation because of different end effects on different bands. The 136-foot end-fed antenna is slightly long for 3.5 Mc., but will work well in the region (3500–3600 kc.) that quadruples into the 14-Mc. band. Bands not listed are not recommended for the particular antenna. The center-fed systems are less critical as to length.

On harmonics, the end-fed and center-fed antennas will not have the same directional characteristics, as explained in the text.

When the same antenna is used for work in several bands, it must be realized that the directional characteristic will vary with the band in use.

Simple Systems

The most practical simple multiband antenna is one that is a half-wavelength long at the lowest frequency and is fed either at the center or one end with an open-wire line. Although the standing-wave ratio on the feedline will not approach 1.0 on any band, if the losses in the line are low the system will be efficient. From the standpoint of reduced feedline radiation, a center-fed system is superior to one that is end-fed, but the end-fed arrangement is often more convenient and should not be ignored as a possibility. The center-fed antenna will not have the same radiation pattern as an end-fed one of the same length, except on frequencies where the over-all length of the antenna is a half-wavelength or less. The end-fed antenna acts like a long-wire antenna on all bands (for which it is longer than a half-wavelength), but the center-fed one acts like two antennas of half that length fed in phase. For example, if a full-wavelength antenna is fed at one end, it will have a radiation pattern as shown in Fig. 10-56, but if it is fed in the center the pattern will be somewhat similar to Fig. 10-48, with the maximum radiation broadside to the wire. Either antenna is a good radiator, but if the radiation pattern is a factor, the point of feed must be considered.

Since multiband operation of an antenna does not permit matching of the feedline, some attention must be paid to the length of the feedline if convenient transmitter-coupling arrangements are to be obtained. Table 10-II gives some suggested antenna and feeder lengths for multiband operation. In general, the length of the feedline should be some integral multiple of a quarter wavelength at the lowest frequency.

Antennas for Restricted Space

If the space available for the antenna is not large enough to accommodate the length necessary for a half-wave at the lowest frequency to be used, quite satisfactory operation can be secured by using a shorter antenna and making up the missing length in the feeder system. The antenna itself may be as short as a quarter wavelength and still radiate fairly well, although of course it will not be as effective as one a half-wave long. Nevertheless, such a system is useful where operation on the desired band otherwise would be impossible.

Resonant feeders are a practical necessity with such an antenna system, and a center-fed antenna will give best all-around performance. With end feed the feeder currents become badly unbalanced.

With center feed practically any convenient length of antenna can be used, if the feeder length is adjusted to accommodate at least

TABLE 10-III
Antenna and Feeder Lengths for Short Multiband Antennas, Center-Fed

Antenna Length (ft.)	*Feeder Length (ft.)*	*Band*	*Type of Tuning*
100	38	3.5 Mc. 7 Mc. 14 Mc. 28 Mc.	parallel series series series or parallel
67.5	34	3.5 Mc. 7 Mc. 14 Mc. 28 Mc.	series parallel parallel parallel
50	43	7 Mc. 14 Mc. 28 Mc.	parallel parallel parallel
33	51	7 Mc. 14 Mc. 28 Mc.	parallel parallel parallel
33	31	7 Mc. 14 Mc. 28 Mc.	parallel series parallel

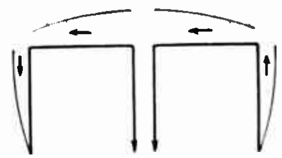


Fig. 10-60 — Folded arrangement for shortened antennas. The total length is a half-wave, not including the feeders. The horizontal part is made as long as convenient and the ends dropped down to make up the required length. The ends may be bent back on themselves like feeders to cancel radiation partially. The horizontal section should be at least a quarter wave long.

one half-wave around the whole system.

A practical antenna of this type can be made as shown in Fig. 10-59. Table 10-III gives a few recommended lengths. However, the antenna can be made any convenient length, provided the total length of wire is a half-wavelength at the lowest frequency, or an integral multiple of a half-wavelength.

Bent Antennas

Since the field strength at a distance is proportional to the current in the antenna, the high-current part of a half-wave antenna (the center quarter wave, approximately) does most of the radiating. Advantage can be taken of this fact when the space available does not permit erecting an antenna a half-wave long. In this case the ends may be bent, either horizontally or vertically, so that the total length equals a half-wave, even though the straightaway horizontal length may be as short as a quarter wave. The operation is illustrated in Fig. 10-60. Such an antenna will be a somewhat better radiator than a quarter-wavelength antenna on the lowest frequency, but is not so desirable for multiband operation because the ends play an increasingly important part as the frequency is raised. The performance of the system in such a case is difficult to predict, especially if the ends are vertical (the most convenient arrangement) because of the complex combination of horizontal and vertical polarization which results as well as the dissimilar directional characteristics. However, the fact that the radiation pattern is incapable of prediction does not detract from the general usefulness of the antenna.

Long-Wire Directive Arrays

● THE "V" ANTENNA

It has been emphasized that, as the antenna length is increased, the lobe of maximum radiation makes a more acute angle with the

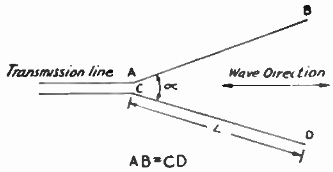


Fig. 10-61 — The basic "V" antenna, made by combining two long wires.

wire. Two such wires may be combined in the form of a horizontal "V" so that the main lobes from each wire will reinforce along a line bisecting the angle between the wires. This increases both gain and directivity, since the lobes in directions other than along the bisector cancel to a greater or lesser extent. The horizontal "V" antenna therefore transmits best in either direction (is bidirectional) along a line bisecting the "V" made by the two wires. The power gain depends upon the length of the wires. Provided the necessary space is available, the "V" is a simple antenna to build and operate. It can also be used on harmonics, so that it is suitable for multiband work. The "V" antenna is shown in Fig. 10-61.

Fig. 10-62 shows the dimensions that should be followed for an optimum design to obtain maximum power gain for different-sized "V" antennas. The longer systems give good performance in multiband operation. Angle ∝ is approximately equal to twice the angle of maximum radiation for a single wire equal in length to one side of the "V."

The wave angle referred to in Fig. 10-62 is the vertical angle of maximum radiation. Tilting the whole horizontal plane of the "V" will tend to increase the low-angle radiation off the low end and decrease it off the high end.

The gain increases with the length of the

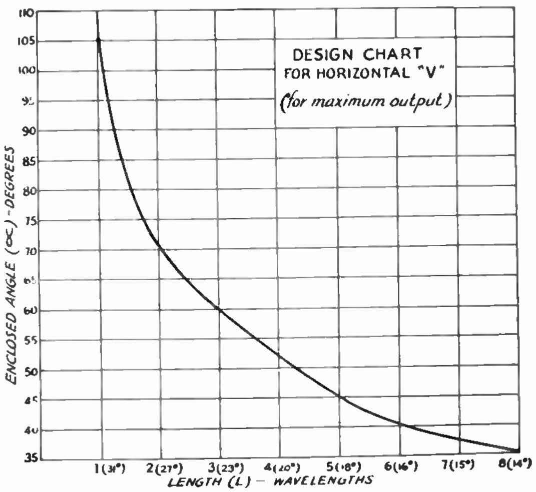


Fig. 10-62 — Design chart for horizontal "V" antennas, giving the enclosed angle between sides *vs.* the length of the wires. Values in parentheses represent approximate wave angle for height of one-half wavelength.

wires, but is not exactly twice the gain for a single long wire as given in Fig. 10-55. In the longer lengths the gain will be somewhat increased, because of mutual coupling between the wires. A "V" eight wavelengths on a leg, for instance, will have a gain of about 12 db. over a half-wave antenna, whereas twice the gain of a single eight-wavelength wire would be only approximately 9 db.

The two wires of the "V" must be fed out of phase, for correct operation. A resonant line may simply be attached to the ends, as shown in Fig. 10-61. Alternatively, a quarter-wave matching section may be employed and the antenna fed through a nonresonant line. If the antenna wires are made multiples of a half-wave in length (use Equation 10-X for computing the length), the matching section will be closed at the free end. A stub can be connected across the resonant line to provide a match, as described later.

THE RHOMBIC ANTENNA

The horizontal rhombic or "diamond" antenna is shown in Fig. 10-63. Like the "V," it requires a great deal of space for erection, but it is capable of giving excellent gain and directivity. It also can be used for multiband operation. In the terminated form shown in Fig. 10-63, it operates like a nonresonant transmission line, without standing waves, and is unidirectional. It may also be used without the terminating resistor, in which case there are standing waves on the wires and the antenna is bidirectional.

The important quantities influencing the design of the rhombic antenna are shown in Fig. 10-63. While several design methods may be used, the one most applicable to the conditions existing in amateur work is the so-called "compromise" method. The chart of Fig. 10-64 gives design information based on a given length and wave angle to determine the remaining optimum dimensions for best operation. Curves for values of length of two, three

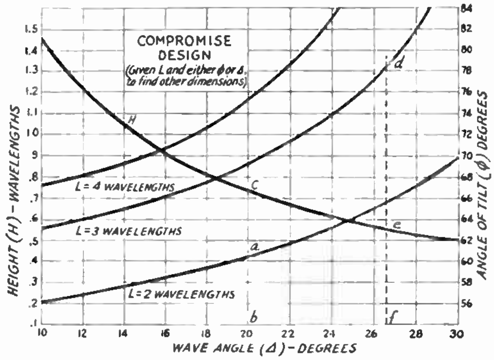


Fig. 10-64 — Compromise-method design chart for rhombic antennas of various leg lengths and wave angles. The following examples illustrate the use of the chart:

(1) Given:

Length (L) = 2 wavelengths
Desired wave angle (Δ) = 20°.

To Find: H, Φ.

Method:

Draw vertical line through point a (L = 2 wavelengths) and point b on abscissa (Δ = 20°). Read angle of tilt (Φ) for point a and height (H) from intersection of line ab at point c on curve H.

Result:

Φ = 60.5°.
H = 0.73 wavelength.

(2) Given:

Length (L) = 3 wavelengths.
Angle of tilt (Φ) = 78°.

To Find: H, Δ.

Method:

Draw a vertical line from point d on curve L = 3 wavelengths at Φ = 78°. Read intersection of this line on curve H (point e) for height, and intersection at point f on the abscissa for Δ.

Result:

H = 0.56 wavelength.
Δ = 26.6°.

and four wavelengths are shown, and any intermediate values may be interpolated.

With all other dimensions correct, an increase in length causes an increase in power gain and a slight reduction in wave angle. An increase in height also causes a reduction in wave angle and an increase in power gain, but not to the same extent as a proportionate increase in length. For multiband work, it is satisfactory to design the rhombic antenna on the basis of 14-Mc. operation, which will permit work from the 7- to 28-Mc. bands as well.

A value of 800 ohms is correct for the terminating resistor for any properly-constructed rhombic, and the system behaves as a pure resistive load under this condition. The terminating resistor must be capable of safely dissipating one-half the power output (to eliminate the rear pattern), and should be noninductive. Such a resistor may be made up from a carbon or graphite rod or from a long 800-ohm transmission line using

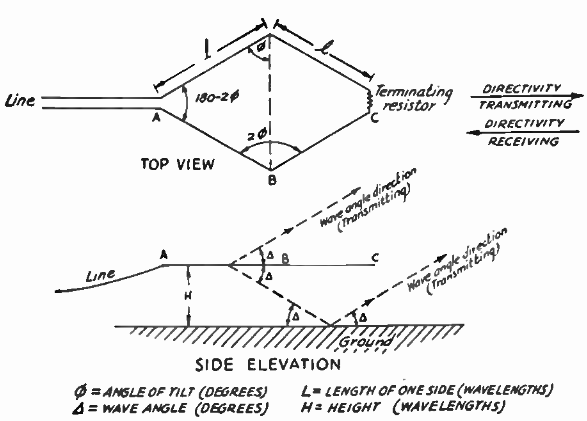


Fig. 10-63 — The horizontal rhombic or diamond antenna, terminated. Important design dimensions are indicated; details in text.

resistance wire. If the carbon rod or a similar form of lumped resistance is used, the device should be suitably protected from weather effects, i.e., it should be covered with a good asphaltic compound and sealed in a small lightweight box or fiber tube. Suitable nonreactive terminating resistors are also available commercially.

For feeding the antenna, the antenna impedance will be matched by an 800-ohm line, which may be constructed from No. 16 wire spaced 20 inches or from No. 18 wire spaced 16 inches. The 800-ohm line is somewhat ungainly to install, however, and may be replaced by an ordinary 600-ohm line with only a negligible mismatch. Alternatively, a matching section may be installed between the antenna terminals and a low-impedance line. However, when such an arrangement is used, it will be necessary to change the matching-section constants for each different band on which operation is contemplated.

The same design details apply to the unterminated rhombic as to the terminated type. When used without a terminating resistor, the system is bidirectional. Resonant feeders are preferable for the unterminated rhombic. A nonresonant line may be used by incorporating a matching section at the antenna, but is not readily adaptable to satisfactory multiband work.

Rhombic antennas will give a power gain of 8 to 12 db. or more for leg lengths of two to four wavelengths, when constructed according to the charts given. In general, the larger the antenna, the greater the power gain.

Directive Arrays with Driven Elements

By combining individual half-wave antennas into an **array** with suitable spacing between the antennas (called **elements**) and feeding power to them simultaneously, it is possible to make the radiated fields from the individual elements add in a favored direction, thus increasing the field strength in that direction as compared to that produced by one antenna element alone. In other directions the fields will more or less oppose each other, giving a reduction in field strength. Thus a power gain in the desired direction is secured at the expense of a power reduction in other directions.

Besides the spacing between elements, the instantaneous direction of current flow (*phase*) in individual elements determines the directivity and power gain. There are several methods of arranging the elements. If they are strung end to end, so that all lie on the same straight line, the elements are said to be **collinear**. If they are parallel and all lying in the same plane, the elements are said to be **broadside** when the phase of the current is the same in all, and **end-fire** when the currents are not in phase. Elements that receive power from the transmitter through the transmission line **are called driven elements.**

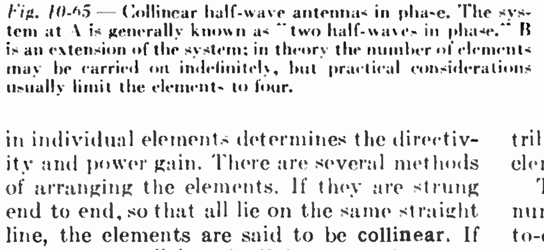


Fig. 10-65 — Collinear half-wave antennas in phase. The system at A is generally known as "two half-waves in phase." B is an extension of the system; in theory the number of elements may be carried on indefinitely, but practical considerations usually limit the elements to four.

The power gain of a directive system increases with the number of elements. The proportionality between gain and number of elements is not simple, however. The gain depends upon the effect that the spacing and phasing has upon the radiation resistance of the elements, as well as upon their number.

Collinear Arrays

Simple forms of collinear arrays, with the current distribution, are shown in Fig. 10-65. The two-element array at A is popularly known as "two half-waves in phase." It will be recognized as simply a center-fed antenna operated at its second harmonic. The way in which the number of elements may be extended for increased directivity and gain is shown in Fig. 10-65B. Note that quarter-wave phasing sections are used between elements; these give the reversal in phase necessary to make the currents in individual antenna elements all flow in the same direction at the same instant. Any phase-reversing section may be used as a quarter-wave matching section for attaching a nonresonant feeder, or a resonant transmission line may be substituted for any of the quarter-wave sections. Also, the antenna may be end-fed by any of the systems previously described, or any element may be center-fed. It is best to feed at the center of the array, so that the energy will be distributed as uniformly as possible among the elements.

The gain and directivity depend upon the number of elements and their spacing, center-to-center. This is shown by Table 10-IV. Although three-quarter wave spacing gives greater gain, it is difficult to construct a suitable phase-reversing system when the ends of the antenna elements are widely separated. For this reason, the half-wave spacing is most generally used in actual practice.

Collinear arrays may be mounted either horizontally or vertically. Horizontal mount-

TABLE 10-IV
Theoretical Gain of Collinear Half-Wave Antennas

Spacing between centers of adjacent half-waves	*Number of half-waves in array vs. gain in db.*				
	2	3	4	5	6
½ wave	1.8	3.3	4.5	5.3	6.2
¾ wave	3.2	4.8	6.0	7.0	7.8

ing gives increased horizontal directivity, while the vertical directivity remains the same as for a single element at the same height. Vertical mounting gives the same horizontal pattern as a single element, but concentrates the radiation at low angles. It is seldom practicable to use more than two elements vertically at frequencies below 14 Mc. because of the excessive height required.

Broadside Arrays

Parallel antenna elements with currents in phase may be combined as shown in Fig. 10-66 to form a **broadside** array, so named because

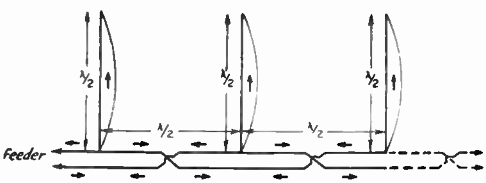


Fig. 10-66 — Broadside array using parallel half-wave elements. Arrows indicate the direction of current flow. Transposition of the feeders is necessary to bring the antenna currents in phase. Any reasonable number of elements may be used. The array is bidirectional, with maximum radiation "broadside" or perpendicular to the antenna plane (perpendicularly through this page).

the direction of maximum radiation is broadside to the plane containing the antennas. Again the gain and directivity depend upon the number of elements and the spacing, the gain for different spacings being shown in Fig. 10-67. Half-wave spacing generally is used, since it simplifies the problem of feeding the system when the array has more than two elements. Table 10-V gives theoretical gain as a function of the number of elements with half-wave spacing.

Broadside arrays may be suspended either with the elements all vertical or with them horizontal and one above the other (**stacked**). In the former case the horizontal pattern becomes quite sharp, while the vertical pattern is the same as that of one element alone. If the array is suspended horizontally, the horizontal pattern is equivalent to that of one element while the vertical pattern is sharpened, giving low-angle radiation.

Broadside arrays may be fed either by resonant transmission lines or through quarter-wave matching sections and nonresonant lines. In Fig. 10-66, note the "crossing over" of the feeders, which is necessary to bring the elements into proper phase relationship.

Combined Broadside and Collinear Arrays

Broadside and collinear arrays may be combined to give both horizontal and vertical directivity, as well as additional gain. The general plan of constructing such antennas is shown in Fig. 10-68. The lower angle of radiation resulting from stacking elements in the vertical plane is desirable at the higher frequencies. In general, doubling the number of elements in an array by stacking will raise the gain from 2 to 4 db., depending upon whether vertical or horizontal elements are used — that is, whether the stacked elements are of the broadside or collinear type.

The arrays in Fig. 10-68 are shown fed from one end, but this is not especially desirable in the case of large arrays. Better distribution of energy between elements, and hence better over-all performance, will result when the feeders are attached as nearly as possible to the center of the array. Thus, in the eight-element array at A, the feeders could be introduced at the middle of the transmission line between the second and third set of elements, in which case the connecting line would not be transposed between the second and third set of elements. Alternatively, the antenna could be constructed with the transpositions as shown and the feeder connected between the adjacent ends of either the second or third pair of collinear elements

A four-element array of the general type shown in Fig. 10-68B, known as the "lazy-H" antenna, has been quite frequently used. This arrangement is shown, with the feed point indicated, in Fig. 10-69.

End-Fire Arrays

Fig. 10-70 shows a pair of parallel half-wave elements with currents out of phase. This is known as an **end-fire** array, because it radiates best along the line of the antennas, as shown.

The end-fire array may be used either ver-

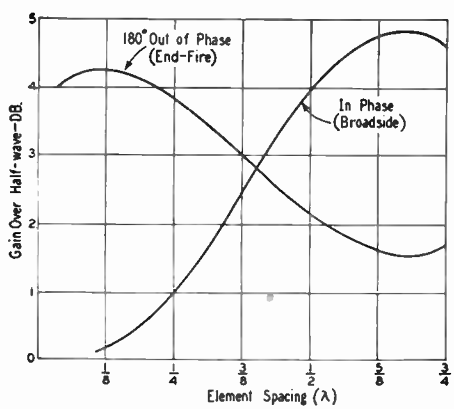


Fig. 10-67 — Gain *vs.* spacing for two parallel half-wave elements combined as either broadside or end-fire arrays.

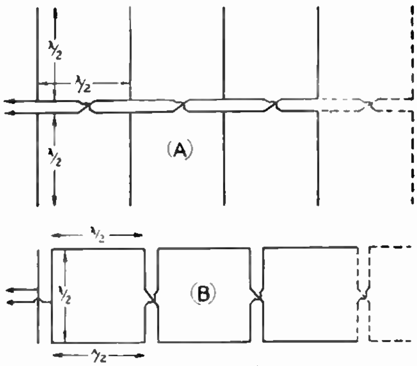


Fig. 10-68 — Combination broadside and collinear arrays. A, with vertical elements; B, with horizontal elements. Both arrays give low-angle radiation. Two or more sections may be used. The gain in db. will be equal, approximately, to the sum of the gain for one set of broadside elements (Table 10-V) plus the gain of one set of collinear elements (Table 10-IV). For example, in A each broadside set has four elements (gain 7 db.) and each collinear set two elements (gain 1.8 db.), giving a total gain of 8.8 db. In B, each broadside set has two elements (gain 4 db.) and each collinear set three elements (gain 3.3 db.), making the total gain 7.3 db. The result is not strictly accurate, because of mutual coupling between the elements, but is good enough for practical purposes.

tically or horizontally (elements at the same height), and is well adapted to amateur work because it gives maximum gain with relatively close element spacing. Fig. 10-67 shows how the gain varies with spacing. End-fire elements may be combined with additional collinear and broadside elements to give a further increase in gain and directivity.

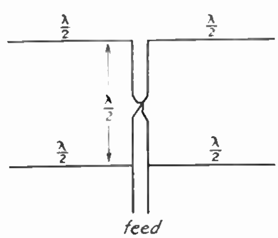


Fig. 10-69 — A four-element combination broadside-collinear array, popularly known as the "lazy-H" antenna. A closed quarter-wave stub may be used at the feed point to match into a 600-ohm transmission line, or resonant feeders may be attached at the point indicated. The gain over a half-wave antenna is 5 to 6 db.

Either resonant or nonresonant lines may be used with this type of array. Nonresonant lines preferably are matched to the antenna through a quarter-wave matching section or phasing stub.

Phasing

Figs. 10-68 and 10-70 illustrate a point in connection with feeding a phased antenna system which sometimes is confusing. In Fig. 10-70, when the transmission line is connected as at A there is no crossover in the line connecting the two antennas, but when the transmission line is connected to the center of the connecting line the crossover becomes necessary (B). This is because in B the two halves of the connecting line are simply *branches* of the same line. In other words, even though the connecting line in B is a half-wave in length, it is not actually a half-wave line but *two quarter-wave lines in parallel.* The same thing is true of the untransposed line of Fig. 10-68B. Note that, under these conditions, the antenna elements are in phase when the line is not transposed, and out of phase when the transposition is made. The opposite is the case when the half-wave line simply joins two antenna elements and does not have the feedline connected to its center, as in Fig. 10-67.

Adjustment of Arrays

With arrays of the types just described, using half-wave spacing between elements, it

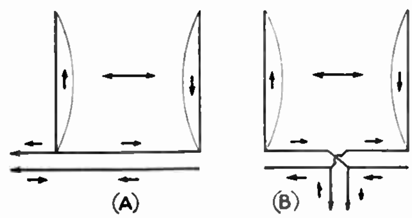


Fig. 10-70 — End-fire arrays using parallel half-wave elements. The elements are shown with half-wave spacing to illustrate feeder connections. In practice, closer spacings are desirable, as shown by Fig. 10-67. Direction of maximum radiation is shown by the large arrows.

will usually suffice to make the length of each element that given by Equations 10-I or 10-J. The half-wave phasing lines between the parallel elements should be of open-wire construction, and their length can be calculated from:

$$\textit{Length of half-wave line} \text{ (feet)} = \frac{480}{\textit{Freq.} \text{ (Mc.)}} \qquad \textbf{(10-O)}$$

Example: A half-wavelength phasing line for 28.8 Mc. would be $\frac{480}{28.8} = 16.66$ feet $= 16$ feet 8 inches.

The spacing between elements can be made equal to the length of the phasing line. No special adjustments of line or element length or spacing are needed, provided the formulas are followed closely.

TABLE 10-V

Theoretical Gain *vs.* Number of Broadside Elements (Half-Wave Spacing)

No. of elements	*Gain*
2	4 db.
3	5.5
4	7
5	8
6	9

With collinear arrays of the type shown in Fig. 10-65B, the same formula may be used for the element length, while the length of the quarter-wave phasing section can be found from the following formula:

$$\textit{Length of quarter-wave line (feet)} = \frac{240}{\textit{Freq. (Mc.)}} \qquad \textbf{(10-P)}$$

Example: A quarter-wavelength phasing line for 14.25 Mc. would be $\frac{240}{14.25} = 16.84$ feet $= 16$ feet 10 inches.

If the array is fed in the center it should not be necessary to make any particular adjustments, although, if desired, the whole system can be resonated by connecting an r.f. ammeter in the shorting link of each phasing section and moving the link back and forth to find the maximum-current position. This refinement is hardly necessary in practice, however, so long as all elements are the same length and the system is symmetrical.

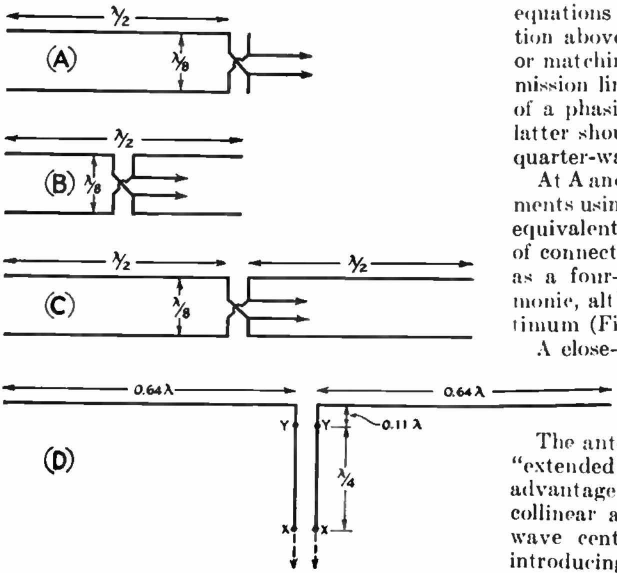


Fig. 10-71 — Simple directive-antenna systems. A is a two-element end-fire array; B is the same array with center feed, which permits use of the array on the second harmonic, where it becomes a four-element array with quarter-wave spacing. C is a four-element end-fire array with ⅛-wave spacing. D is a simple two-element broadside array using extended in-phase antennas ("extended double-Zepp"). The gain of A and B is slightly over 4 db. On the second harmonic, B will give about 5-db. gain. With C, the gain is approximately 6 db., and with D, approximately 3 db. In A, B and C, the phasing line contributes about 1/16 wavelength to the transmission line; when B is used on the second harmonic, this contribution is ⅛ wavelength. Alternatively, the antenna ends may be bent to meet the transmission line, in which case each feeder is simply connected to one antenna. In D, points Y-Y indicate a quarter-wave point (high current) and X-X a half-wave point (high voltage). The line may be extended in multiples of quarter waves if resonant feeders are to be used. A, B and C may be suspended on wooden spreaders. The plane containing the wires should be parallel to the ground.

The phasing sections can be made of 300-ohm Twin-Lead, if low power is used. However, the lengths of the phasing sections must be only 84 per cent of the length obtained in the two formulas above.

Example: The half-wavelength line for 28.8 Mc. would become $0.84 \times 16.66 = 13.99$ feet $=$ 14 feet 0 inches

Using Twin-Lead for the phasing sections is most useful in arrays such as that of Fig. 10-65B, or any other system in which the element spacing is not controlled by the length of the phasing section.

Simple Arrays

Several simple directive-antenna systems using driven elements have achieved rather wide use among amateurs. Four of these systems are shown in Fig. 10-71. Tuned feeders are assumed in all cases; however, a matching section readily can be substituted if a non-resonant transmission line is preferred. Dimensions given are in terms of wavelength; actual lengths can be calculated from the equations for the antenna and from the equation above for the resonant transmission line or matching section. In cases where the transmission line proper connects to the midpoint of a phasing line, only *half* the length of the latter should be added to the line to find the quarter-wave point.

At A and B are two-element end-fire arrangements using close spacing. They are electrically equivalent; the only difference is in the method of connecting the feeders. B may also be used as a four-element array on the second harmonic, although the spacing is not quite optimum (Fig. 10-67) for such operation.

A close-spaced four-element array is shown at C. It will give about 2 db. more gain than the two-element array.

The antenna at D, commonly known as the "extended double-Zepp," is designed to take advantage of the greater gain possible with collinear antennas having greater than half-wave center-to-center spacing, but without introducing feed complications. The elements are made longer than a half-wave in order to bring this about. The gain is 3 db. over a single half-wave antenna, and the broadside directivity is fairly sharp.

The antennas of A and B may be mounted either horizontally or vertically; horizontal suspension (with the elements in a plane parallel to the ground) is recommended, since this tends to give low-angle radiation without an unduly sharp horizontal pattern. Thus these systems are useful for coverage over a wide horizontal angle. The system at C, when mounted horizontally, will have a sharper horizontal pattern than the two-element arrays because of the effect of the collinear arrangement. The vertical pattern, however, will be the same as that of the antennas in A and B.

Antennas for 160 Meters

Results on 1.8 Mc. will depend to a large extent on the antenna system and the time of day or night. Almost any random long wire that can be tuned to resonance will work

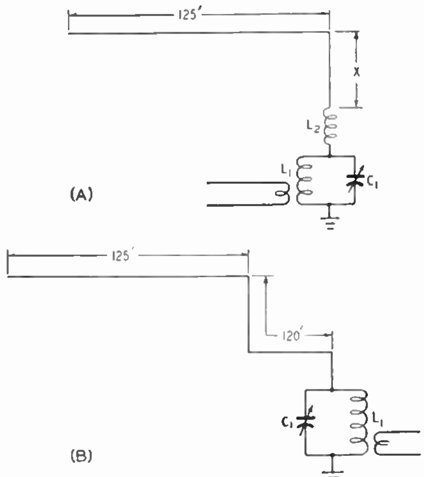


Fig. 10-72 — Bent antenna for the 160-meter band. In the system at A, the vertical portion (length X) should be made as long as possible. In either antenna system, L_1C_1 should resonate at 1900 kc., roughly. To adjust L_2 in antenna A, resonate L_1C_1 alone to the operating frequency, then connect it to the antenna system and adjust L_2 for maximum loading. Further loading can be obtained by increasing the coupling between L_1 and the link.

during the night but it will generally be found very ineffective during the day. A vertical antenna — or rather an antenna from which the radiation is predominantly vertically polarized — is probably the best for 1.8-Mc. operation. A horizontal antenna (horizontally polarized radiation) will give better results during the night than the day because daytime absorption in the ionosphere is so high at this frequency that the reflected wave is too weak to be useful. At night the performance improves because nighttime ionosphere conditions generally permit the reflected wave to return to earth without too much attenuation. The vertically-polarized radiator gives a strong ground wave that is effective day or night, and it is to be preferred on 1.8 Mc.

There is another reason why a vertical antenna is better than a horizontal for 160-meter operation. The low-angle radiation from a horizontal antenna ⅛ or ¼ wavelength above ground is almost insignificant. Any reasonable height is small in terms of wavelength, so that a horizontal antenna on 160 meters is a poor radiator at angles useful for long distances ("long", that is, for this band). Its chief usefulness is over relatively short distances at night.

Bent Antennas

Since ideal vertical antennas are generally out of the question for practical amateur work, the best compromise is to bend the antenna in such a way that the high-current portions of the antenna run vertically. It is, of course, advisable to place the antenna so that the highest currents in the antenna occur at the highest points above actual ground. Two antenna systems designed along these lines are shown in Fig. 10-72. The antenna at A uses a loading coil, L_2, to increase the electrical length of the antenna to a half wavelength, so that the antenna can be fed at its high-voltage point through the coupling circuit L_1C_1. The antenna of Fig. 10-72B uses a full half-wavelength of wire but is bent so that the high-current portion runs vertically. The horizontal portion running to L_1C_1 should run 8 or 10 feet above ground.

Grounds

A good ground connection is generally important on 160 meters. The ideal system is a number of wire radials buried a foot or two underground and extending 50 to 100 feet from the central connection point. As many radials as possible should be used.

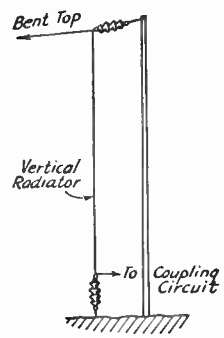


Fig. 10-73 — An arrangement for keeping the main radiating portion of the antenna vertical.

If the soil is good (not rocky or sandy) and generally moist, a low-resistance connection to the cold-water pipe system in the house will often serve as an adequate ground system. The connection should be made close to where the pipe enters the ground, and the surface of the pipe should be scraped clean before tightening the ground clamp around the pipe.

A 6- or 8-foot length of 1-inch water pipe, driven into the soil at a point where there is considerable natural moisture, can be used for the ground connection. Three or four pipes, driven into the ground 8 or 10 feet apart and all joined together at the top with heavy wire, are more effective than the single pipe.

The use of a counterpoise is recommended where a buried system is not practicable or where a pipe ground cannot be made to have low resistance because of poor soil conditions. A counterpoise consists of a number of wires supported from 6 to 10 feet above the surface of the ground. Generally the wires are spaced 10 to 15 feet apart and located to form a square or polygonal configuration under the vertical portion of the antenna.

Matching the Antenna to the Line

Except in the several cases of half-wave antennas mentioned earlier, most antenna systems do not have center impedances that readily match open-wire lines or available solid-dielectric ones. However, any antenna can be matched to practically any line by any of the several means to be described. The matching is accomplished by first resonating the antenna to the proper frequency and then introducing either a matching transformer between the antenna and the line or by applying corrective stubs to the line.

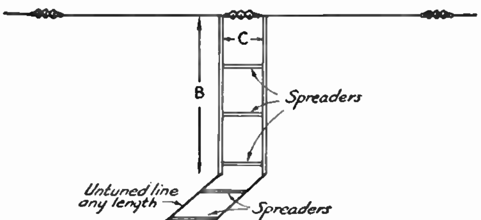


Fig. 10-74 — The "Q" antenna, using a quarter-wave impedance-matching section with close-spaced conductors.

An impedance mismatch of 10 or 20 per cent is of little consequence so far as power transfer to the antenna is concerned. It is relatively easy to get the standing-wave ratio down to 1.5- or 2-to-1, a perfectly satisfactory condition in practice. Of considerably greater importance is the necessity for getting the currents in the two wires balanced, both as to amplitude and phase. If the currents are not the same at corresponding points on adjacent wires and the loops and nodes do not also occur at corresponding points, there will be considerable radiation loss. Perfect balance can be brought about only by perfect symmetry in the line, particularly with respect to ground. This symmetry should extend to the coupling apparatus at the transmitter.

In the following discussion of ways in which different types of lines may be matched to the antenna, a half-wave antenna is used as an example. Other types of antennas may be treated by the same methods, making due allowance for the order of impedance that appears at the end of the line when more elaborate systems are used.

"Q"-Section Transformer

The impedance of a two-wire line of ordinary construction (400 to 600 ohms) can be matched to the impedance of the center of a half-wave antenna by utilizing the impedance-transforming properties of a quarter-wave line, Equation 10-B. The matching section must have low surge impedance and therefore is commonly constructed of large-diameter conductors such as aluminum or copper tubing, with fairly-close spacing. This system is known as the "Q" antenna. It is shown in Fig. 10-74. Important dimensions are the length of the antenna itself, the length of the matching section, *B*, the spacing between the two conductors of the matching section, *C*, and the impedance of the untuned transmission line connected to the lower end of the matching section.

The required characteristic impedance for the matching section is

$$Z_m = \sqrt{Z_1 Z_2} \qquad \text{(10-B)}$$

where Z_1 and Z_2 are the antenna and feedline impedances.

Example: To match a 600-ohm line to an antenna presenting a 72-ohm load, the quarter-wave matching section would require a characteristic impedance of $\sqrt{72 \times 600} = \sqrt{43{,}200} = 208$ ohms.

The spacings between conductors of various sizes of tubing and wire for different surge impedances are given in graphical form in Fig. 10-10. With ½-inch tubing, the spacing should be 1.5 inches for an impedance of 208 ohms.

The length of the matching section, *B*, should be equal to a quarter wavelength, and is given by Equation 10-G. The length of the antenna can be calculated from Equations 10-I or 10-J.

This system has the advantage of the simplicity of adjustment of the 75-ohm feeder system and at the same time the superior insulation of an open-wire system.

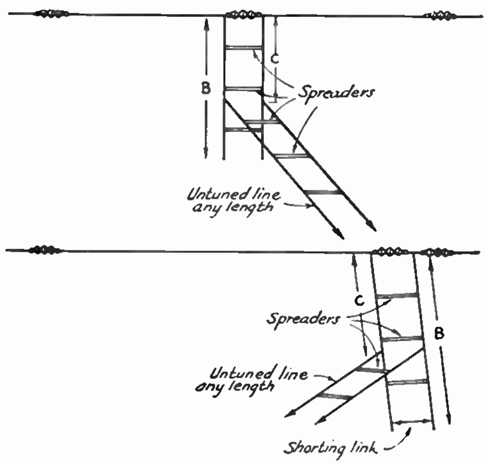


Fig. 10-75 — Antenna systems with quarter-wave open-wire linear impedance-matching transformers.

Linear Transformers

Fig. 10-75 shows two methods of coupling a nonresonant line to an antenna through a quarter-wave linear transformer or matching section. In the case of the center-fed antenna, the free end of the matching section, *B*, is open (high impedance) if the other end is connected to a low-impedance point (current loop) on the antenna. With the end-fed antenna, the free end of the matching section is closed through a shorting bar or link; this end of the section has low impedance, since the other end is connected to a high-impedance point on the antenna.

When the connection between the matching section and the antenna is unbalanced, as in the end-fed system, it is important that the antenna be the right length for the operating frequency if a good match is to be obtained. The balanced center-fed system is less critical in this respect. The shorting-bar method of tuning the center-fed system to resonance may be used if the matching section is extended to a half-wavelength, bringing a current loop at the free end.

In the center-fed system, the antenna and matching section should be cut to lengths found from Equations 10-I, 10-X and 10-P. Any necessary on-the-ground adjustment can be made by adding to or clipping off the open ends of the matching section. In the end-fed system the matching section can be adjusted

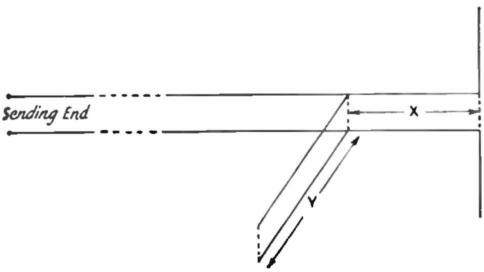


Fig. 10-76 — When antenna and transmission line differ in impedance, they may be matched by a short length of transmission line, Y, called a stub. Determination of the critical dimensions, X and Y, for proper matching depends on whether the stub is open or closed at the end.

by making the line a little longer than necessary and adjusting the system to resonance by moving the shorting link up and down. Resonance can be determined by exciting the antenna at the proper frequency from a temporary antenna near by and measuring the current in the shorting bar by a low-range r.f. ammeter or galvanometer using one of the devices of this type described in the chapter on measurements. The position of the bar should be adjusted for maximum current reading. This should be done before the transmission line is attached to the matching section.

The position of the line taps will depend upon the impedance of the line as well as on the antenna impedance at the point of connection. The procedure is to take a trial point, apply power to the transmitter, and then check the transmission line for standing waves. This can be done by measuring the current in, or voltage along, the wires. At any one position along the line the currents in the two wires should be identical. Readings taken at intervals of a

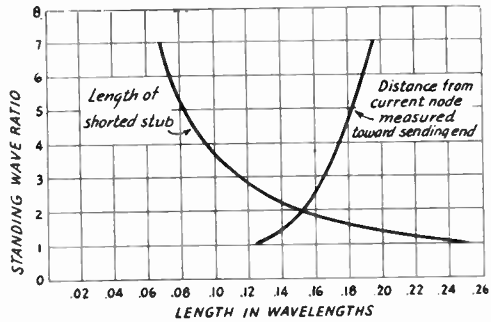


Fig. 10-77 — Graph for determining position and length of a *shorted* stub. Dimensions may be converted to linear units after values have been taken from the graph.

quarter wavelength will indicate whether or not standing waves are present.

It will not usually be possible to obtain complete elimination of standing waves when the matching stub is exactly resonant, but the line taps should be adjusted for the smallest obtainable standing-wave ratio. Then a further "touching up" of the matching-stub tuning will eliminate the remaining standing waves, provided the adjustments are carefully made. The stub must be readjusted, because when resonant it exhibits some reactance as well as resistance at all points except at the ends, and a slight lengthening or shortening of the stub is necessary to tune out this reactance.

Matching Stubs

The operation of the quarter-wave matching transformer of Fig. 10-75 may be considered from another — and more general — viewpoint. Suppose that section C is looked upon simply as a continuation of the transmission line. Then the "free" end of the transformer becomes a "stub" line, shunting a section of the main transmission line. From this viewpoint, matching the line to the antenna becomes a matter of selecting the right type and length of stub and attaching it to the proper spot along the line.

Referring to Fig. 10-76, at any distance (X) from the antenna, the line will have an imped-

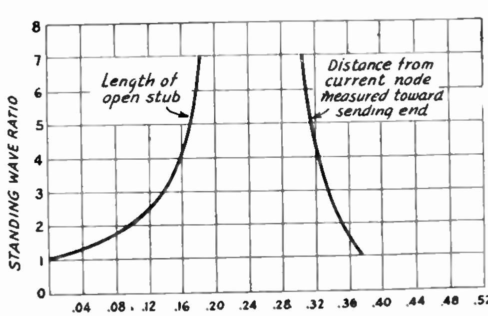


Fig. 10-78 — Graph for determining position and length of an *open* stub. Dimensions may be converted to linear units after values have been taken from the graph.

ance that may be considered to be made up of reactance (either inductive or capacitive) and resistance, in parallel. The reactive component can be eliminated by shunting the line at distance X from the antenna with another reactance equal in value but opposite in sign to the reactance presented by the line at that point. If distance X is such that the line presents an inductive reactance, a corresponding shunting capacitive reactance will be required.

The required compensating reactance may be supplied by shunting the line with a stub cut to proper length, Y. With the reactances canceled only a pure resistance remains as a termination for the remainder of the line between the sending end and the stub, and this resistance can be adjusted to match the characteristic impedance of the line by adjusting the distance X.

Distances X and Y may be determined experimentally, but since their values are interdependent the cut-and-try method is somewhat laborious. If the standing-wave ratio and the positions of the current loops and nodes can be measured, the length and position of the stub can be found from Figs. 10-77 and 10-78.

While it is relatively easy to locate the position of the current (or voltage) loops and nodes by examining the line with a neon bulb, r.f. galvanometer, or pick-up loop and crystal detector, other means are more direct for determining the standing-wave ratio. Several devices of this type are described in Chapter Sixteen, and the use of these also affords a simple method for determining the location of current loops (voltage nodes). With the meter or indicator in the line near the transmitter, points will be found on the transmission line where touching the line with a screwdriver will have a minimum effect on the meter indication. These points correspond to voltage nodes.

Once the standing-wave ratio is known, the length and position of the stub, in terms of wavelength, can be found directly from Figs. 10-77 and 10-78. The wavelength in feet for any frequency can be found from Equation 10-O.

Measuring Standing Waves

In adjusting a "Q-match" or linear transformer, or a delta or "T"-match to an antenna, one of the standing-wave indicators described in Chapter Sixteen should be used. If 300-ohm Twin-Lead is used, the simple "twin-lamp" indicator is the most convenient and the simplest to use. For lines of other impedance, or for coaxial line, the Micro-Match type or the bridge type should be used. In any event, the absolute value of standing-wave ratio is not as important as the proper adjustment for a minimum ratio, since ratios of 1.5-to-1 or less represent good amateur practice.

Where two-wire lines are used, the standing-wave-ratio indicator should give the same reading regardless of the polarity of the transmission line — any discrepancy indicates an unbalance in the line.

Directive Arrays with Parasitic Elements

Parasitic Excitation

The antenna arrays previously described are bidirectional; that is, they will radiate in directions both to the "front" and to the "back" of the antenna system. If radiation is wanted in only one direction, it is necessary to use different element arrangements. In most of these arrangements the additional elements receive power by induction or radiation from the driven element, generally called the "antenna," and reradiate it in the proper phase relationship to achieve the desired effect. These elements are called *parasitic* elements, as contrasted to the driven elements which receive power directly from the transmitter through the transmission line. They are widely used to give additional gain and directivity to simple antennas.

The parasitic element is called a **director** when it reinforces radiation on a line pointing to it from the antenna, and a **reflector** when the reverse is the case. Whether the parasitic element is a director or reflector depends upon the parasitic-element tuning (which usually is adjusted by changing its length) and, particularly when the element is self-resonant, upon the spacing between it and the antenna.

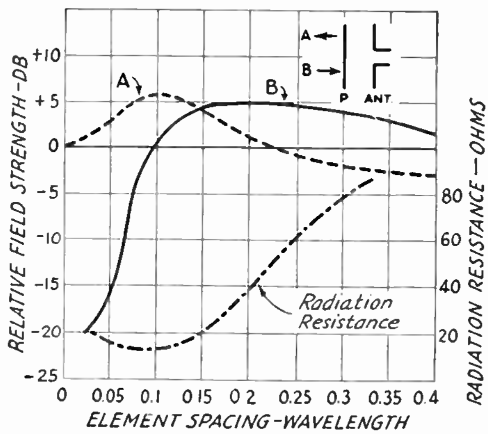


Fig. 10-79 — **Gain *vs.* element spacing for an antenna and one parasitic element. The reference point, 0 db., is the field strength from a half-wave antenna alone. The greatest gain is in direction *A* at spacings of less than 0.14 wavelength, and in direction *B* at greater spacings. The front-to-back ratio is the difference in db. between curves *A* and *B*. Variation in radiation resistance of the driven element also is shown. These curves are for a self-resonant parasitic element. At most spacings the gain as a reflector can be increased by slight lengthening of the parasitic element; the gain as a director can be increased by shortening. This also improves the front-to-back ratio.**

Gain vs. Spacing

The gain of an antenna-reflector or an antenna-director combination varies chiefly with the spacing between the elements. The way in which gain varies with spacing is shown in Fig. 10-79, for the special case of self-resonant parasitic elements. This chart also shows how the attenuation to the "rear" varies with spacing. The same spacing does not necessarily give both maximum forward gain and maximum backward attenuation. Backward attenuation is desirable when the antenna is used for receiving, since it greatly reduces interference coming from the opposite direction to the desired signal.

Element Lengths

The antenna length is given by the formula for a half-wavelength antenna. The director and reflector lengths must be determined experimentally for maximum performance. The preferable method is to aim the antenna at a receiver a mile or more distant and have an observer check the signal strength (on the receiver S-meter) while the reflector or director is adjusted a few inches at a time, until the length which gives maximum signal is found. The attenuation may be similarly checked, the length being adjusted for minimum signal. In general, for best front-to-back ratio the length of a director will be about 4 per cent less than that of the antenna. The reflector will be about 5 per cent longer than the antenna.

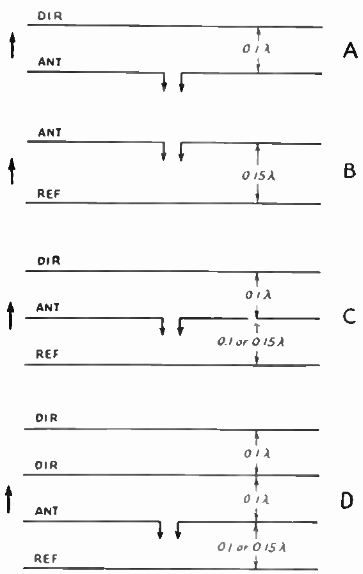


Fig. 10-80 — Half-wave antennas with parasitic elements. A, with director; B, with reflector; C, with both director and reflector; D, two directors and one reflector. Gain is approximately as shown by Fig. 10-79, in the first two cases, and depends upon the spacing and length of the parasitic element. In the three- and four-element arrays a reflector spacing of 0.15 wavelength will give slightly more gain than 0.1-wavelength spacing. Arrows show the direction of maximum radiation.

Simple Systems: the Rotary Beam

Four practical combinations of antenna, reflector and director elements are shown in Fig. 10-80. Spacings which give maximum gain or maximum front-to-back ratio (ratio of power radiated in the desired direction to power radiated in the opposite direction) may be taken from Fig. 10-79. In the chart, the front-to-back ratio in db. will be the sum of gain and attenuation at the same spacing.

Systems of this type are popular for rotary-beam antennas, where the entire antenna system is rotated, to permit its gain and directivity to be utilized for any compass direction. They may be mounted either horizontally (with the plane containing the elements parallel to the earth) or vertically.

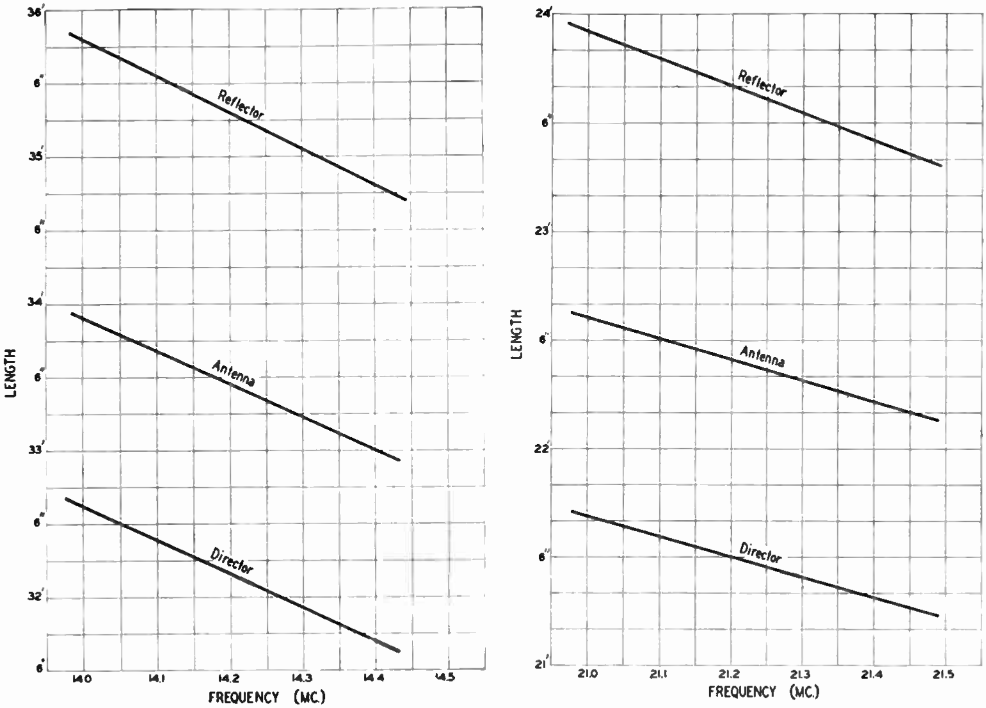
Arrays using more than one parasitic element, such as those shown at C and D in Fig. 10-80, will give more gain and directivity than is indicated for a single reflector or director by the curves of Fig. 10-79. The gain with a properly-adjusted three-element array (antenna, director and reflector) will be 5 to 7 db. over a half-wave antenna. Somewhat higher gain still can be secured by adding a second director to the system, making a four-element array. The front-to-back ratio is correspondingly improved as the number of elements is increased.

The elements in close-spaced (less than one-quarter wavelength element spacing) arrays preferably should be made of tubing of one-half to one-inch diameter. A conductor of large diameter not only has less ohmic resistance but also has lower *Q;* both these factors are important in close-spaced arrays because the impedance of the driven element usually is quite low compared to that of a single half-wave dipole. With 3- and 4-element arrays the radiation resistance of the driven element may be as low as 6 or 8 ohms, so that ohmic losses in the conductor can consume an appreciable fraction of the power. Low radiation resistance means that the antenna will work over only a small frequency range without retuning unless large-diameter conductors are used. In addition, the antenna elements should be rigid because if they are free to move with respect to each other, the array will tend to show troublesome detuning effects under windy conditions.

Feeding Close-Spaced Arrays

While any of the usual methods of feed may be applied to the driven element of a parasitic array, the fact that, with close spacing, the radiation resistance as measured at the center of the driven element drops to a very low value makes some systems more desirable than others. The preferred methods are shown in Fig. 10-82. Resonant feeders are not recommended for lengths greater than a half-wavelength.

The quarter- or half-wave matching stubs shown at A and B in Fig. 10-82 preferably



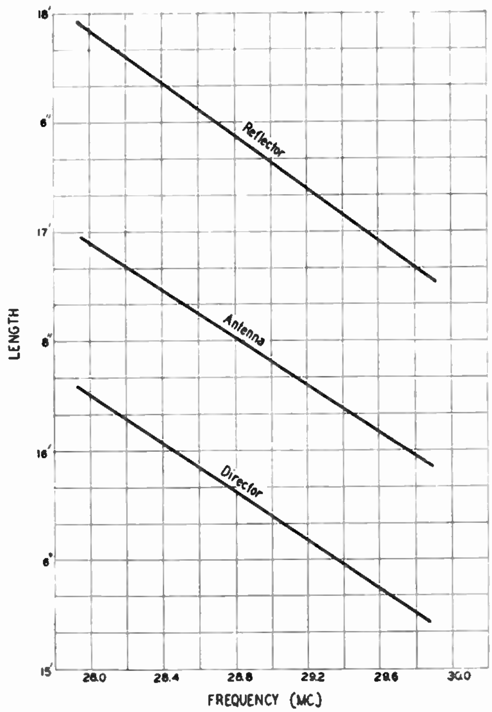


Fig. 10-81 — Director, antenna and reflector lengths for three-element beams, for element spacing of 0.1 to 0.2 wavelength. The greater spacing will result in slightly higher gain. The lengths indicated are for maximum gain — some improvement in front-to-back ratio may be obtained by adjustment of the reflector length.

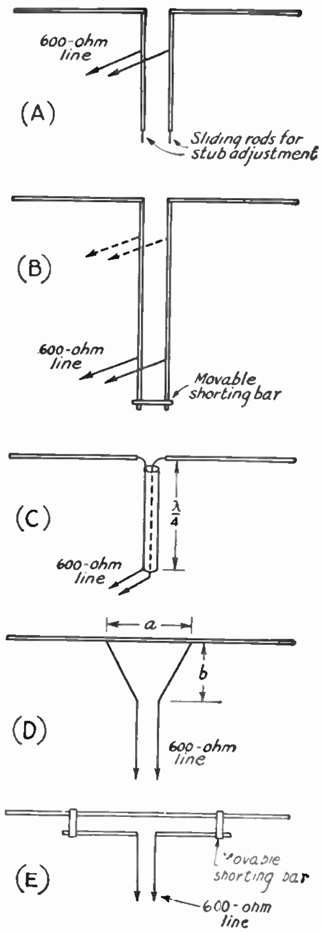


Fig. 10-82 — Recommended methods of feeding the driven antenna element in close-spaced parasitic arrays. The parasitic elements are not shown. A, quarter-wave open stub; B, half-wave closed stub; C, concentric-line quarter-wave matching section; D, delta matching transformer; E, "T" matching transformer. Adjustment details are discussed in the text.

should be constructed of tubing with rather close spacing, in the manner of the "Q" section. This lowers the impedance of the matching section and makes the position of the line taps somewhat less difficult to determine accurately. The line adjustment should be made only with the parasitic elements in place, and after the correct element lengths have been determined it should be checked to compensate for changes likely to occur because of element tuning.

The concentric-line matching section at C will work with fair accuracy into a close-spaced parasitic array of 2, 3 or 4 elements without necessity for adjustment. The line is used as an impedance-inverting transformer, and, if its characteristic impedance is 70 ohms (RG-11/U), it will give a good match to a 600-ohm line when the resistance at the termination is about 8.5 ohms. Over a range of 5 to 15 ohms the mismatch, and therefore the standing-wave ratio, will be less than 2-to-1. The length of the quarter-wave section may be calculated from Equation 10-G.

The delta matching transformer shown at D is probably easier to install, mechanically, than any of the others. The positions of the taps (dimension *a*) must be determined experimentally, along with the length, *b*, by checking the standing-wave ratio on the line as adjustments are made. Dimension *b* should be about 15 per cent longer than *a*.

The system shown at E ("T"-match) resembles the delta match in principles of operation. It has the advantage that, with close spacing between the two parallel conductors, line radiation from the matching section is negligible whereas radiation from a delta may be considerable. It is adjusted by moving the shorting bars, keeping them equidistant from the center, until there are no standing waves on the line. The matching section may be made of the same type of conductor used for the driven element and spaced a few inches from it.

The "folded-dipole" type of antenna may be used as the driven element of a close-spaced parasitic array to secure an impedance step-up to the transmission line and also to broaden the resonance curve of the antenna. The folded dipole consists of two or more half-wave antennas connected together at the ends with the feeder connected to the center of only one of the antennas. The spacing between the parallel antennas should be small — of the order of the spacing used between wires of a transmission line. The current in the system divides in approximate proportion to the areas of the conductors, resulting in an impedance step-up at the input terminals. With two similar conductors (equal areas) the impedance step-up is 4-to-1; if there are three similar conductors (or if the one not connected to the transmission line has twice the diameter of the other) the step-up is 9-to-1; if the ratio of the areas is 3-to-1 the step-up is 16-to-1, and so on. Thus if a 3-conductor dipole (all conductors the same diameter) is used as the driven element of a four-element parasitic array the center impedance of approximately 8 ohms is multiplied by 9 and appears as approximately 72 ohms at the input terminals. Such a system therefore can be fed directly from a 70-ohm line with no additional means for matching.

Fig. 10-83 shows the impedance step-up obtained in a folded dipole when conductors of different sizes are used.

Sharpness of Resonance

Peak performance of a multielement parasitic array depends upon proper phasing or tuning of the elements, which can be exact for one frequency only. In the case of close-spaced arrays, which because of the low radiation resistance usually are quite sharp-tuning, the frequency range over which optimum results can be secured is only of the order of 1 or 2

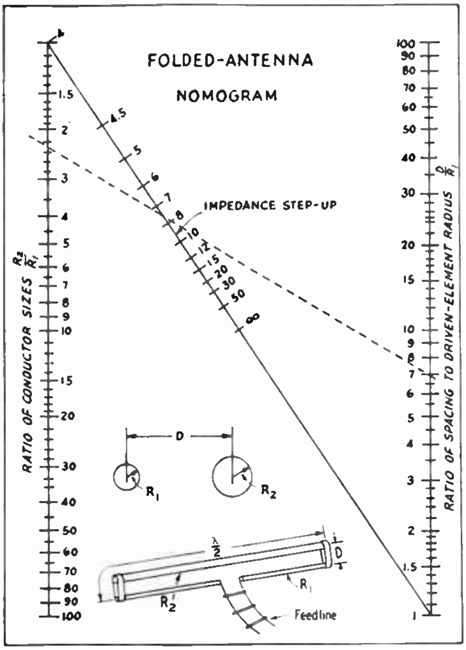


Fig. 10-83 — **Nomogram for computing impedance step-up in a folded dipole with dissimilar conductors. The line at the left is the ratio of conductor diameters, and the line at the right is the ratio of conductor spacing (center-to-center) to the driven-element radius. The solid slanting line is the impedance step-up ratio. Laying a straightedge between any two known quantities will give the value of the third.**

Example: Find the diameter of the large conductor when the driven-element diameter is 0.5 inch, line impedance 300 ohms, antenna impedance 40 ohms, and spacing 1.75 inches.
Impedance step-up required = 300/40 = 7.5
Spacing-to-element-radius ratio = 1.75/0.25 = 7
Laying a straightedge across the figure (dashed line), ratio of conductor diameters = 2.3
Diameter of large conductor = 2.3 × 0.5 = 1.15 inches

per cent of the resonant frequency, or up to about 500 kc. at 28 Mc. However, the antenna can be made to work satisfactorily over a wider frequency range by adjusting the director or directors to give maximum gain at the *highest* frequency to be covered, and by adjusting the reflector to give optimum gain at the *lowest* frequency. This sacrifices some gain at all frequencies, but maintains more uniform gain over a wider frequency range.

As mentioned in the preceding paragraphs, the use of large-diameter conductors will broaden the response curve of an array because the larger diameter lowers the Q. This causes the reactances of the elements to change rather slowly with frequency, with the result that the tuning stays near the optimum over a considerably-wider frequency range than is the case with wire conductors.

Combination Arrays

It is possible to combine parasitic elements with driven elements to form arrays composed of collinear driven and parasitic elements and combination broadside-collinear-parasitic elements. Thus two or more collinear elements might be provided with a collinear reflector or director set, one parasitic element to each driven element. Or both directors and reflectors might be used. A broadside-collinear array could be treated in the same fashion.

When combination arrays are built up, a rough approximation of the gain to be expected may be obtained by adding the gains for each type of combination. Thus the gain of two broadside sets of four collinear arrays with a set of reflectors, one behind each element, at quarter-wave spacing for the parasitic elements, would be estimated as follows: From Table 10-IV, the gain of four collinear elements is 4.5 db. with half-wave spacing; from Fig. 10-67 or Table 10-V, the gain of two broadside elements at half-wave spacing is 4.0 db.; from Fig. 10-79, the gain of a parasitic reflector at quarter-wave spacing is 4.5 db. The total gain is then the sum, or 13 db. for the sixteen elements. Note that using two *sets* of elements in broadside is equivalent to using two elements, so far as gain is concerned; similarly with sets of reflectors, as against one antenna and one reflector. The actual gain of the combination array will depend, in practice, upon the way in which the power is distributed between the various elements and upon the effect which mutual coupling between elements has upon the radiation resistance of the array, and may be somewhat higher or lower than the estimate.

A great many directive-antenna combinations can be worked out by combining elements according to these principles.

RECEIVING ANTENNAS

Nearly all of the properties possessed by an antenna as a radiator also apply when it is used for reception. Current and voltage distribution, impedance, resistance and directional characteristics are the same in a receiving antenna as if it were used as a transmitting antenna. This reciprocal behavior makes possible the design of a receiving antenna of optimum performance based on the same considerations that have been discussed for transmitting antennas.

The simplest receiving antenna is a wire of random length. The longer the wire, the more energy it abstracts from the wave. Because of the high sensitivity of modern receivers, a large antenna is not necessary for picking up signals at good strength. An indoor wire only 15 to 20 feet long will serve at frequencies below the v.h.f. range, although a longer wire outdoors is better.

The use of a tuned antenna improves the operation of the receiver, however, because the signal strength is raised more in proportion to the stray noises picked up than is the case with wires of random length. Since the transmitting antenna usually is given the best loca-

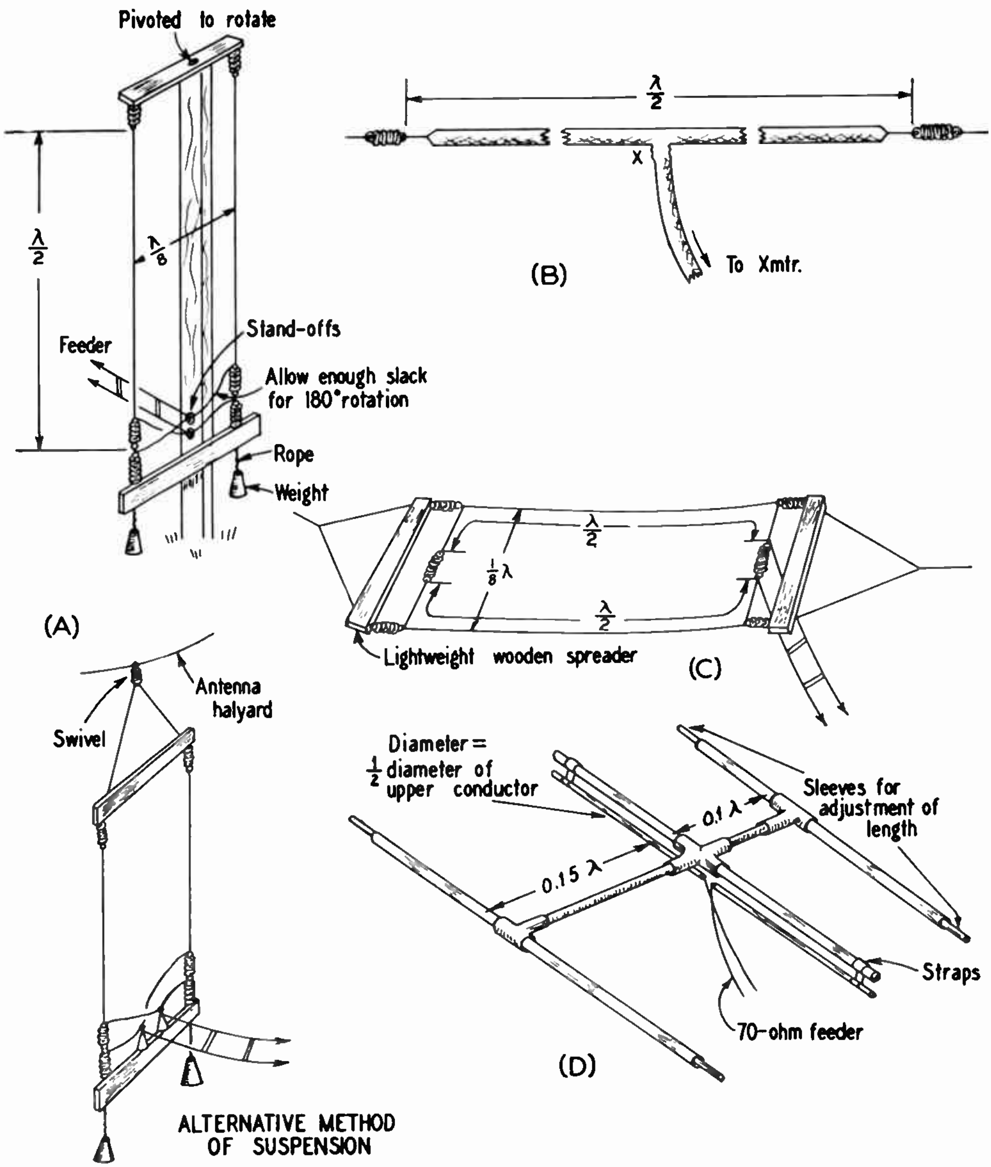
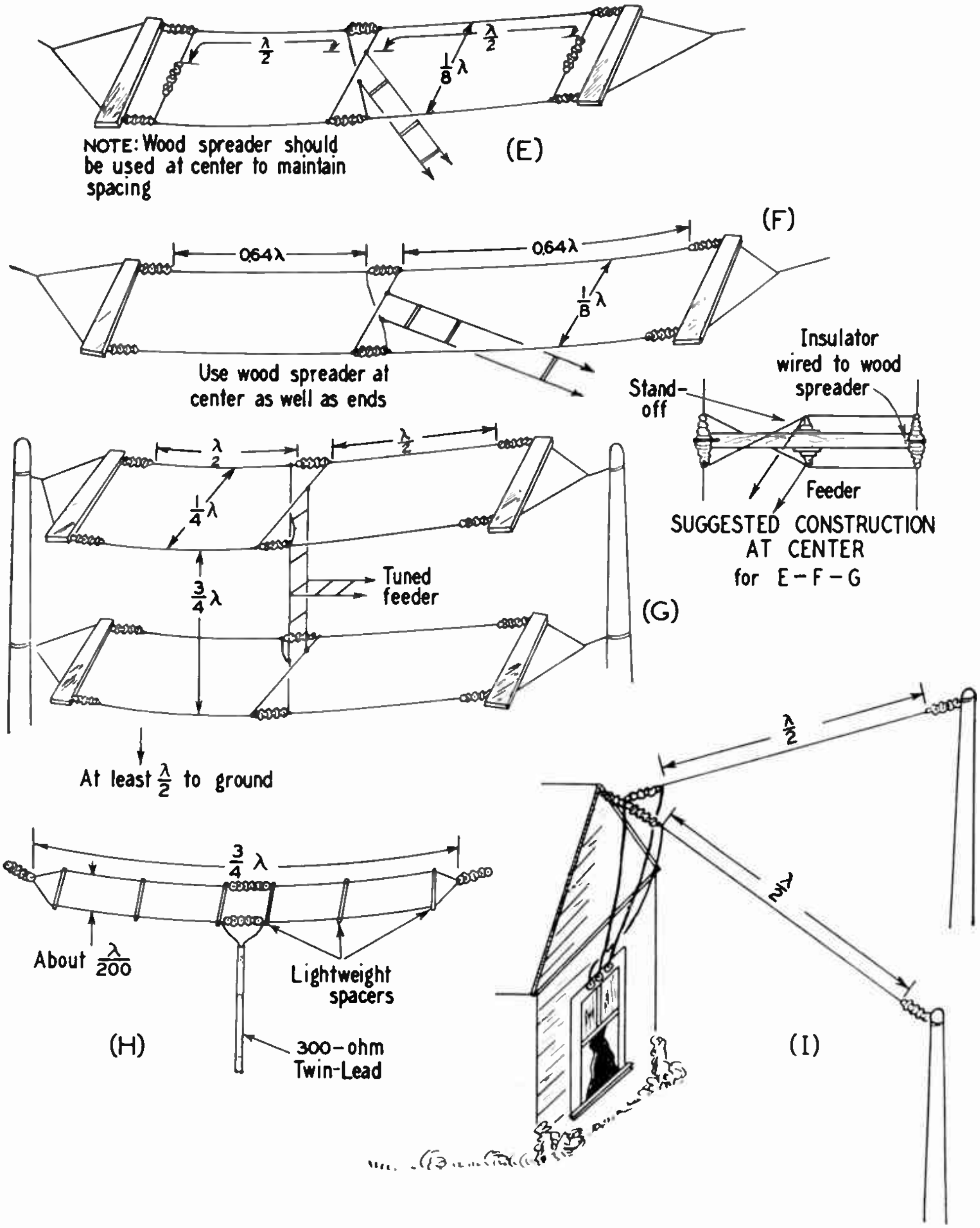


Fig. 10-84 — Some suggested antenna systems. A — Simple bidirectional rotatable end-fire array using ⅛-wave spacing between out-of-phase elements. It is suitable for either 14 or 28 Mc. and can be hand-rotated. It can also be suspended from the halyard holding another antenna, as suggested in the lower drawing. B — Folded dipole using 300-ohm Twin-Lead for both antenna and feeder. The junction X at the center is made by opening one conductor of the antenna section and soldering to the feeder leads. The joint may be made mechanically firm by heating the dielectric with a soldering iron, using extra bits of dielectric for a good bond. C — An end-fire array for use where space is limited. The ends of the two half-wave elements are folded to meet at an insulator in the center. The antenna may be made still shorter by increasing the spacing: spacings up to ¼ wavelength may be used. D — Pipe-assembly three-element beam ("plumber's delight") with folded-dipole driven element. Because all three elements are at the same r.f. potential at their centers it is possible to join them electrically as well as mechanically with no effect on the performance. Provision is made for adjusting the element lengths for optimum performance at a given frequency. E — An extension of the folding principle shown in C. The collinear in-phase elements give additional gain and directivity. F — End-



fire array with extended double-Zepps. This antenna should give a gain of about 7 db. in the direction perpendicular to the line of the antenna. G — An 8-element array combining broadside, end-fire and collinear elements. The gain of an antenna of this type is about 10 db. This antenna also can be used at half the frequency for which it is designed. H — A three-quarter wavelength folded antenna matches 500- or 600-ohm open-wire line, but 300-ohm Twin-Lead will be satisfactory. Its pattern is quite similar to a half-wavelength antenna. Note that, unlike the half-wavelength folded dipole, the far side is *open* at the center. I — Using two half-wave antennas at right angles to change direction. With the three feeders indicated, either antenna alone can be fed as a Zepp and will radiate best perpendicular to its direction. By feeding the two together, leaving the third feeder wire idle, the optimum direction is the bisector of the angle between the wires. This system is most useful at high frequencies.

In these drawings, wavelength dimensions on conductors refer to lengths calculated for the conductor size as described in Equation 10-J.

The feeders to the various directive systems in A, C, E, F and G must be tuned if used as shown. For one-band operation, matching stubs may be attached to the feeders if a matched line is desired.

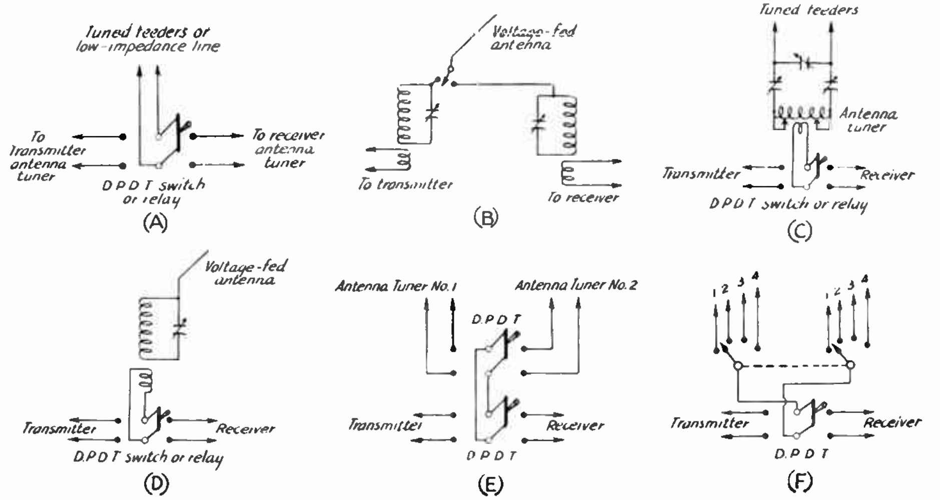


Fig. 10-85 — **Antenna-switching arrangements for various types of antennas and coupling systems. A — For tuned lines with separate antenna tuners or low-impedance lines. B — For a voltage-fed antenna. C — For a tuned line with a single antenna tuner. D — For a voltage-fed antenna with a single tuner. E — For two tuned-line antennas with a tuner for each antenna or for two low-impedance lines. F — For combinations of several two-wire lines.**

tion, it can also be expected to serve best for receiving. This is especially true when a directive antenna is used, since the directional effects and power gain of directive transmitting antennas are the same for receiving as for transmitting.

In selecting a directional receiving antenna it is preferable to choose a type that gives very little response in all but the desired direction (small minor lobes). This is even more important than high gain in the desired direction, because the cumulative response to noise and unwanted-signal interference in the smaller lobes may offset the advantage of increased desired-signal gain. The feedline from the antenna should be balanced so that it will not pick up signals and destroy the directivity.

Antenna Switching

Switching of the antenna from receiver to transmitter is commonly done with a change-over relay, connected in the antenna leads or the coupling link from the antenna tuner. If the relay is one with a 115-volt a.c. coil, the switch or relay that controls the transmitter plate power will also control the antenna relay. If the convenience of a relay is not desired, porcelain knife switches can be used and thrown by hand.

Typical arrangements are shown in Fig. 10-85. If coaxial line is used, the use of a coaxial relay is recommended, although on the lower-frequency bands a regular switch or change-over relay will work almost as well.

Antenna Construction

The use of good materials in the antenna system is important since the antenna is exposed to wind and weather. To keep electrical losses low, the wires in the antenna and feeder system must have good conductivity and the insulators must have low dielectric loss and surface leakage, particularly when wet.

For short antennas, No. 14 gauge hard-drawn enameled copper wire is a satisfactory conductor. For long antennas and directive arrays, No. 14 or No. 12 enameled copper-clad steel wire should be used. It is best to make feeders and matching stubs of ordinary soft-drawn No. 14 or No. 12 enameled copper wire, since hard-drawn or copper-clad steel wire is difficult to handle unless it is under considerable tension at all times. The wires should be all in one piece; where a joint cannot be avoided, it should be carefully soldered.

In building a resonant two-wire feeder, the spacer insulation should be of as good quality as in the antenna insulators proper. For this reason, good ceramic spacers are advisable. Wooden dowels boiled in paraffin may be used with untuned lines, but their use is not recommended for tuned lines. The wooden dowels can be attached to the feeder wires by drilling small holes and binding them to the feeders with wire.

At points of maximum voltage, insulation is most important, and Pyrex glass, Isolantite or steatite insulators with long leakage paths are recommended for the antenna. Glazed porcelain also is satisfactory. Insulators should be cleaned once or twice a year, especially if they are subjected to much smoke and soot.

In most cases poles or masts are desirable to lift the antenna clear of surrounding buildings, although in some locations the antenna will be sufficiently in the clear when strung

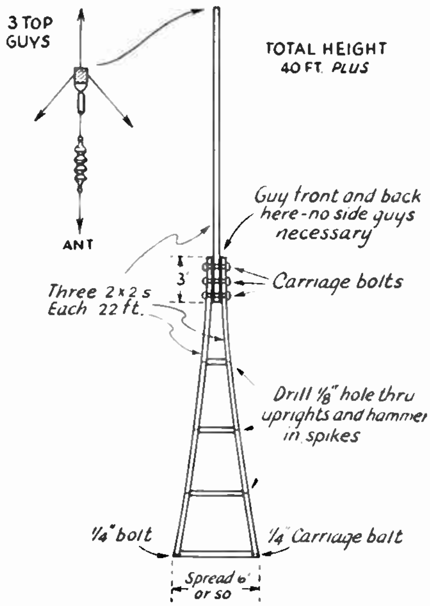


Fig. 10-86 — **Details of a simple 40-foot "A"-frame mast suitable for erection in locations where space is limited.**

from one chimney to another or from a chimney to a tree. Small trees usually are not satisfactory as points of suspension for the antenna because of their movement in windy weather. If the antenna is strung from a point near the center of the trunk of a large tree, this difficulty is not so serious. Where the antenna wire must be strung from one of the smaller branches, it is best to tie a pulley firmly to the branch and run a rope through the pulley to the antenna, with the other end of the rope attached to a counterweight near the ground. The counterweight will keep the tension on the antenna wire reasonably constant even when the branches sway or the rope tightens and stretches with varying climatic conditions.

"A"-FRAME MAST

The simple and inexpensive mast shown in Fig. 10-86 is satisfactory for heights up to 35 or 40 feet. Clear, sound lumber should be selected. The completed mast may be protected by two or three coats of house paint.

If the mast is to be erected on the ground, a couple of stakes should be driven to keep the bottom from slipping and it may then be "walked up" by a pair of helpers. If it is to go on a roof, first stand it up against the side of the building and then hoist it from the roof, keeping it vertical. The whole assembly is light enough for two men to perform the complete operation — lifting the mast, carrying it to its permanent berth, and fastening the guys — with the mast vertical all the while. It is entirely practicable, therefore, to erect this type of mast on any small, flat area of roof.

By using 2 × 3s or 2 × 4s, the height may be extended up to about 50 feet. The 2 × 2 is too flexible to be satisfactory at such heights.

SIMPLE 40-FOOT MAST

The mast shown in Fig. 10-87 is relatively strong, easy to construct, readily dismantled, and costs very little. Like the "A"-frame, it is suitable for heights of the order of 40 feet.

The top section is a single 2 × 3, bolted at the bottom between a pair of 2 × 3s with an overlap of about two feet. The lower section thus has two legs spaced the width of the narrow side of a 2 × 3. At the bottom the two legs are bolted to a length of 2 × 4 which is set in the ground. A short length of 2 × 3 is placed between the two legs about halfway up the bottom section, to maintain the spacing.

The two back guys at the top pull against the antenna, while the three lower guys prevent buckling at the center of the pole.

The 2 × 4 section should be set in the ground so that it faces the proper direction, and then made vertical by lining it up with a plumb bob. The holes for the bolts should be drilled beforehand. With the lower section laid on the ground, bolt *A* should be slipped in place through the three pieces of wood and tightened just enough so that the section can turn freely on the bolt. Then the top section may be bolted in place and the mast pushed up, using a ladder or another 20-foot 2 × 3 for the job. As the mast goes up, the slack in the guys can be taken up so that the whole structure is in some meas-

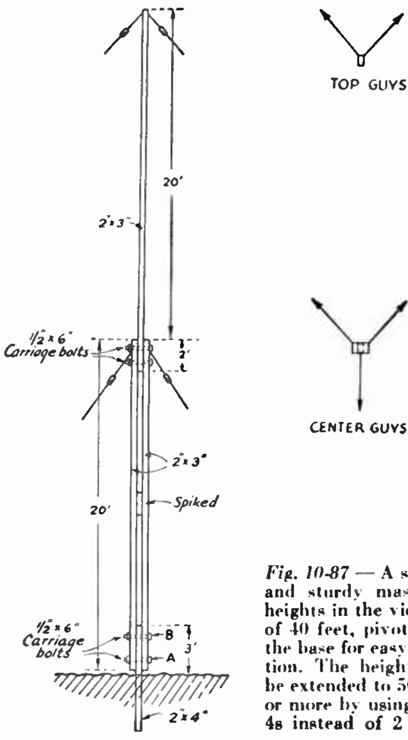


Fig. 10-87 — **A simple and sturdy mast for heights in the vicinity of 40 feet, pivoted at the base for easy erection. The height can be extended to 50 feet or more by using 2 × 4s instead of 2 × 3s.**

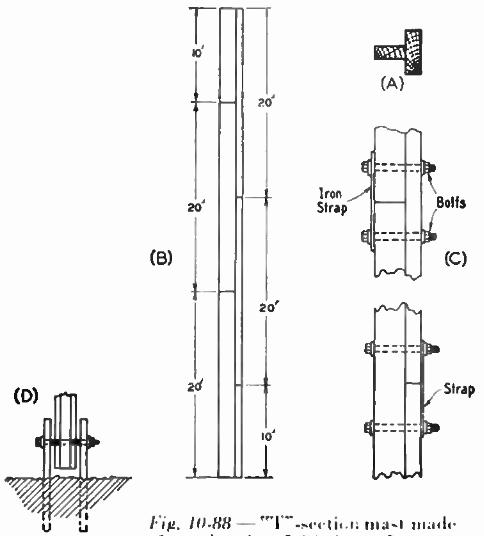


Fig. 10-88 — "T"-section mast made of overlapping 2 × 4s or 2 × 6s.

ure continually supported. When the mast is vertical, bolt *B* should be slipped in place and both *A* and *B* tightened. The lower guys can then be given a final tightening, leaving those at the top a little slack until the antenna is pulled up, when they should be adjusted to pull the top section into line.

"T"-SECTION MAST

A type of mast suitable for heights up to about 80 feet is shown in Fig. 10-88. The mast is built up by butting 2 × 4 or 2 × 6 timbers flatwise against a second 2 × 4, as shown at A, with alternating joints in the edgewise and flatwise sections. The construction can be carried out to greater lengths simply by continuing the 20-foot sections. Longer or shorter sections may be used.

The method of making the joints is shown at C. Quarter-inch or 3/16-inch iron, 1½ to 2 inches wide, is recommended for the straps, with ½-inch bolts to hold the pieces together. One bolt should be run through the pieces midway between joints, to provide additional rigidity.

Although there are many ways in which such a mast can be secured at the base, the "cradle" illustrated at D has many advantages. Heavy timbers set firmly in the ground, spaced far enough apart so the base of the mast will pass between them, hold a large carriage bolt or steel bar which serves as a bearing. The bolt goes through a hole in the mast so that it is pivoted at the bottom.

Half of the guys can be tightened up before the mast leaves the ground by using four sets of guys, one in front, one directly in the rear, and one on each side at right angles to the direction the mast will face. The mast should be guyed every twenty feet and at the top, at each of the joints in the edgewise sections, the guy wires being wrapped around the pole for added strength.

For heights up to 50 feet, 2 × 4-inch members may be used throughout. For greater heights, use 2 × 4s for the edgewise sections; 2 × 6-inch pieces will do for the flat sections.

POLE AND TOWER SUPPORTS

Poles, which often may be purchased at a reasonable price from the local telephone or power company, have the advantage that they do not require guying unless they are called upon to carry a very heavy load. The life of a pole can be extended many years by proper precautions before erecting, and regular maintenance thereafter.

Before setting the pole, it should be given four or five coats of creosote, applying it liberally so it can soak into and preserve the wood. The bottom of the pole and the part that will be buried in the ground should have a generous coating of hot pitch, poured on while the pole is warm. This will keep termites out and prevent rotting.

The pole should be set in the ground four to eight feet depending upon the height. It is a good idea to pour concrete around the bottom three feet of the base, packing the rest of the excavation with soil. The concrete will help hold the pole against strong winds. After filling the hole with dirt, a stream from a hose should be played on the dirt slowly for several hours.

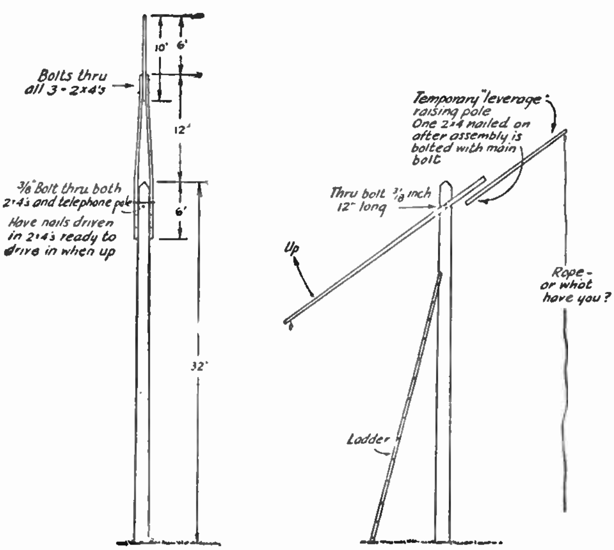


Fig. 10-89 — This type of mast may be carried to a height of fifty feet or more. No guy wires are required.

This will help to settle the soil quickly.

If desired, the pole may be extended by the arrangement shown in Fig. 10-89. Three 2 × 4s are required for the top section, two being 18 feet long and one 10 feet long. The 10-foot section is placed between the other two and bolted in place. A half-inch hole should be bored through the pole about 2 feet from its top and through both 18-foot 2 × 4s about 5 feet from their bottom ends, which are spread apart to fit the top of the pole. The bottom end of the extension is then hauled up to the top of the pole and bolted loosely so that the section can be swung up into place by the leverage of another 2 × 4 temporarily fastened to the section, as shown in Fig. 10-89.

Lattice towers built of wood should be assembled with brass screws and casein glue, rather than with nails which work loose in a short time. A tower constructed in this manner will give trouble-free service if treated with a coat of paint every year.

In painting outside structures, use pure white lead, thinned with three parts of pure linseed oil to one part of turpentine, for the first coat on new wood. The use of a drier is not recommended if the paint will possibly dry without it, since it may cause the paint to peel after a short time. For the second and third coats pure white lead thinned only with pure

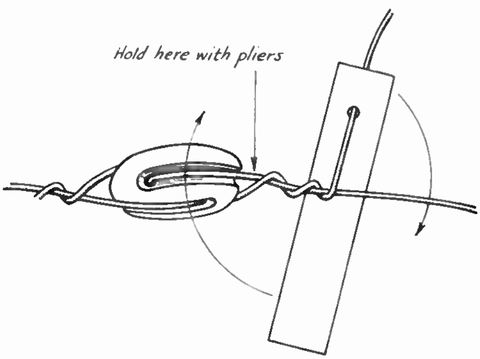


Fig. 10-90 — Using a lever for twisting heavy guy wires.

linseed oil is recommended. Plenty of time for drying should be allowed between coats. White paint will last fifty per cent longer than any colored paint.

● GUYS AND GUY ANCHORS

For masts or poles up to about 50 feet, No. 12 iron wire is a satisfactory guy-wire material. Heavier wire or stranded cable may be used for taller poles or poles installed in locations where the wind velocity is high.

More than three guy wires in any one set usually are unnecessary. If a horizontal antenna is to be supported, two guy wires in the top set will be sufficient in most cases. These should run to the rear of the mast about 100 degrees apart to offset the pull of the antenna.

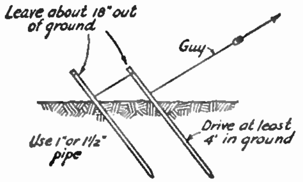


Fig. 10-91 — Pipe guy anchors. One pipe is sufficient for small masts, but two installed as shown will provide the additional strength required for the larger poles.

Intermediate guys should be used in sets of three, one running in a direction opposite to that of the antenna, while the other two are spaced 120 degrees either side. This leaves a clear space under the antenna. The guy wires should be adjusted to pull the pole slightly back from vertical before the antenna is hoisted so that when the antenna is pulled up tight the mast will be straight.

When raising a mast that is big enough to tax the facilities available, it is some advantage to know nearly exactly the length of the guys. Those on the side on which the pole is lying can then be fastened temporarily to the anchors beforehand, which assures that when the pole is raised, those holding opposite guys will be able to pull it into nearly-vertical position with no danger of its getting out of control. The guy lengths can be figured by the right-angled-triangle rule that "the sum of the squares of the two sides is equal to the square of the hypotenuse." In other words, the distance from the base of the pole to the anchor should be measured and squared. To this should be added the square of the pole length to the point where the guy is fastened. The square root of this sum will be the length of the guy.

Guy wires should be broken up by strain insulators, to avoid the possibility of resonance at the transmitting frequency. Common practice is to insert an insulator near the top of each guy, within a few feet of the pole, and then cut each section of wire between the insulators to a length which will not be resonant either on the fundamental or harmonics. An insulator every 25 feet will be satisfactory for frequencies up to 30 Mc. The insulators should be of the "egg" type with the insulating material under compression, so that the guy will not part if the insulator breaks.

Twisting guy wires onto "egg" insulators may be a tedious job if the guy wires are long and of large gauge. The simple time- and finger-saving device shown in Fig. 10-90 can be made from a piece of heavy iron or steel by drilling a hole about twice the diameter of the guy wire about a half inch from one end of the piece. The wire is passed through the insulator, given a single turn by hand, and then held with a pair of pliers at the point shown in the sketch. By passing the wire through the hole in the iron and rotating the iron as shown, the wire may be quickly and neatly twisted.

Guy wires may be anchored to a tree or building when they happen to be in convenient spots. For small poles, a 6-foot length of 1-inch pipe driven into the ground at an angle will suffice. Additional bracing will be provided by using two pipes, as shown in Fig. 10-91.

HALYARDS AND PULLEYS

Halyards or ropes and pulleys are important items in the antenna-supporting system. Particular attention should be directed toward the choice of a pulley and halyards for a high mast since replacement, once the mast is in position, may be a major undertaking if not entirely impossible.

Galvanized-iron pulleys will have a life of only a year or so. Especially for coastal-area installations, marine-type pulleys with hardwood blocks and bronze wheels and bearings should be used.

An arrangement that has certain advantages over a pulley when a mast is used is shown in Fig. 10-92. In case the rope breaks, it may be possible to replace it by heaving a line over the brass rod, making it unnecessary to climb or lower the pole.

For short antennas and temporary installations, heavy clothesline or window-sash cord may be used. However, for more permanent jobs, ⅜-inch or ½-inch waterproof hemp rope should be used. Even this should be replaced about once a year to insure against breakage.

Nylon rope, used during the war as glider tow rope, is, of course, one of the best materials for halyards, since it is weatherproof and has extremely long life.

It is advisable to carry the pulley rope back up to the top in "endless" fashion in the manner of a flag hoist so that if the antenna breaks close to the pole, there will be a means for pulling the hoisting rope back down.

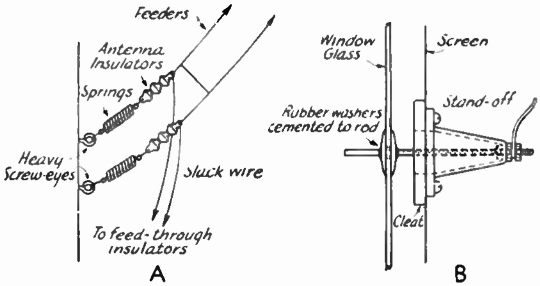


Fig. 10-93 — A — Anchoring feeders takes the strain from feed-through insulators or window glass. B — Going through a full-length screen, a cleat is fastened to the frame of the screen on the inside. Clearance holes are cut in the cleat and also in the screen.

BRINGING THE ANTENNA OR FEEDLINE INTO THE STATION

The antenna or transmission line should be anchored to the outside wall of the building, as shown in Fig. 10-93, to remove strain from the lead-in insulators. Holes cut through the walls of the building and fitted with feed-through insulators are undoubtedly the best means of bringing the line into the station. The holes should have plenty of air clearance about the conducting rod, especially when using tuned lines that develop high voltages. Probably the best place to go through the walls is the trimming board at the top or bottom of a window frame which provides flat surfaces for lead-in insulators. Either cement or rubber gaskets may be used to waterproof the exposed joints.

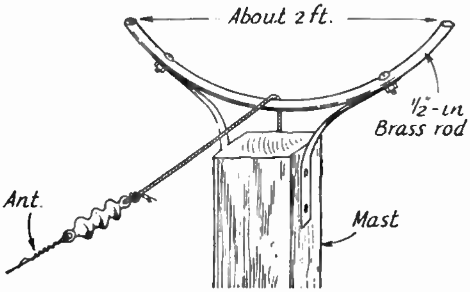


Fig. 10-92 — This device is much easier than a pulley to "rethread" when the rope breaks.

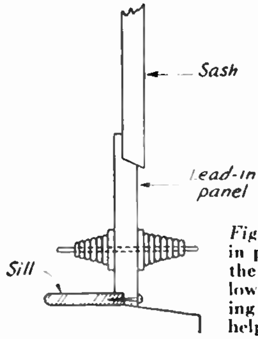


Fig. 10-94 — An antenna lead-in panel may be placed over the top sash or under the lower sash of a window. Sealing the overlapping joint will help make it weatherproof.

Where such a procedure is not permissible, the window itself usually offers the best opportunity. One satisfactory method is to drill holes in the glass near the top of the upper sash. If the glass is replaced by plate glass, a stronger job will result. Plate glass may be obtained from automobile junk yards and drilled before placing in the frame. The glass itself provides insulation and the transmission line may be fastened to bolts fitting the holes. Rubber gaskets will render the holes waterproof. The lower sash should be provided with stops to prevent damage when it is raised. If the window has a full-length screen, the scheme shown in Fig. 10-93B may be used.

As a less permanent method, the window may be raised from the bottom or lowered from the top to permit insertion of a board which carries the feed-through insulators. This lead-in arrangement can be made weatherproof by making an overlapping joint between the board and window sash, as shown in Fig. 10-94, and

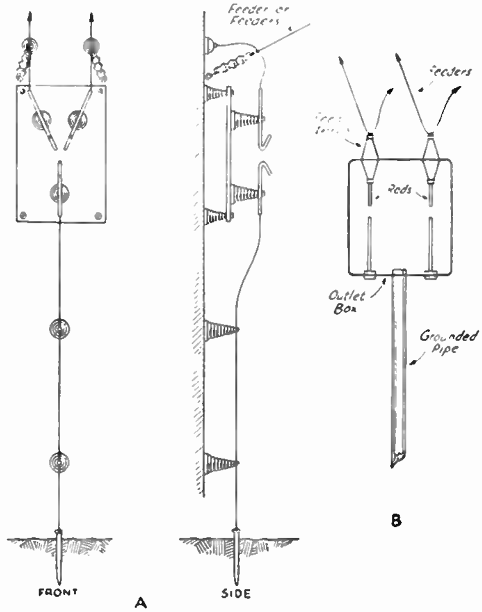


Fig. 10-95 — Low-loss lightning arresters for transmitting-antenna installations.

covering the opening between sashes with a sheet of soft rubber from a discarded inner tube.

● LIGHTNING PROTECTION

An ungrounded radio antenna, particularly if large and well elevated, is a lightning hazard. When grounded, it provides a measure of protection. Therefore, grounding switches or lightning arresters should be provided. Examples of construction of low-loss arresters are shown in Fig. 10-95. At A, the arrester electrodes are mounted by means of stand-off insulators on a fireproof asbestos board. At B, the electrodes are enclosed in a standard steel outlet box. The gaps should be made as small as possible without danger of breakdown during operation. Lightning-arrester systems require the best ground connection obtainable.

The most positive protection is to ground the antenna system when it is not in use; grounded flexible wires provided with clips for connection to the feeder wires may be used. The ground lead should be short and run, if possible, directly to a driven pipe or water pipe where it enters the ground outside the building.

Rotary-Beam Construction

It is a distinct advantage to be able to shift the direction of a beam antenna at will, thus securing the benefits of power gain and directivity in any desired compass direction. A favorite method of doing this is to construct the antenna so that it can be rotated in the horizontal plane. Obviously, the use of such rotatable antennas is limited to the higher frequencies — 14 Mc. and above — and to the simpler antenna-element combinations if the structure size is to be kept within practicable bounds. For the 14- and 28-Mc. bands such antennas usually consist of two to four elements and are of the parasitic-array type described earlier in this chapter. At 50 Mc. and higher it becomes possible to use more elaborate arrays because of the shorter wavelength and thus obtain still higher gain. Antennas for these bands are described in Chapter Fourteen.

The problems in rotary-beam construction are those of providing a suitable mechanical support for the antenna elements, furnishing a means of rotation, and attaching the transmission line so that it does not interfere with the rotation of the system.

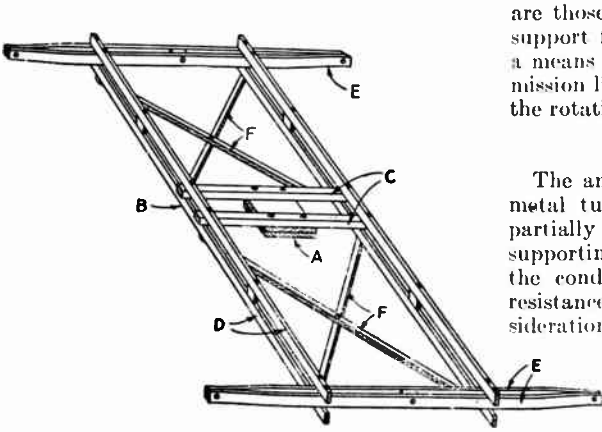


Fig. 10-96 — Easily-built supporting structure for horizontal rotary beams. Made chiefly of 1 × 2″ wood strip, it is strong yet lightweight. Antenna elements are supported on stand-off insulators on the arms, *E*. The length of the *D* sections will depend upon the element spacing, while the length of the *E* sections and the spacing between the *D* sections should be ¼ to ½ the length of the antenna elements.

Elements

The antenna elements usually are made of metal tubing so that they will be at least partially self-supporting, thus simplifying the supporting structure. The large diameter of the conductor is beneficial also in reducing resistance, which becomes an important consideration when close-spaced elements are used. Dural tubes often are used for the elements, and thin-walled corrugated steel tubes with copper coating also are available for this purpose. The elements frequently are constructed of sections of telescoping tubing making length adjustments for tuning quite easy. Electrician's thin-walled conduit also is suitable for rotary-beam elements.

If steel elements are used, special precautions

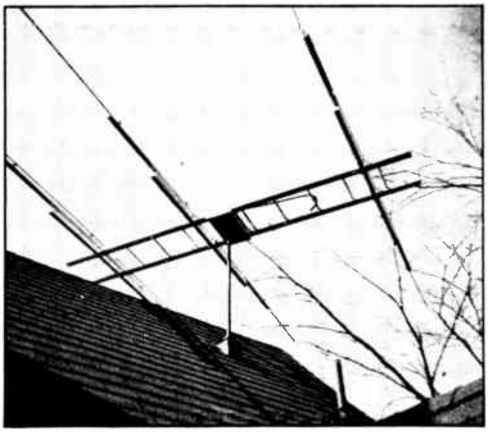


Fig. 10-97 — **A ladder-supported 3-element 28-Mc. beam. It is mounted on a pipe mast that projects through a bearing in the roof and is turned from the attic operating room. (W1MRK in August, 1946, *QST*.)**

should be taken to prevent rusting. Even copper-coated steel does not stand up indefinitely, since the coating usually is too thin. The elements should be coated both inside and out with slow-drying aluminum paint. For coating the inside, a spray gun may be used, or the paint may be poured in one end while rotating the tubing. The excess paint may be caught as it comes out the bottom end and poured through again until it is certain that the entire inside wall has been covered. The ends should then be plugged up with corks sealed with glyptal varnish.

Supports

The supporting framework for a rotary beam usually is made of wood but sometimes of metal, using as lightweight construction as is consistent with the required strength. Generally, the frame is not required to hold much weight, but it must be extensive enough so that the antenna elements can be supported near enough to their ends to prevent excessive sag, and it must have sufficient strength to stand up under the maximum wind in the locality. The design of the frame will depend chiefly on the size of the antenna elements, whether they are mounted horizontally or vertically, and the method to be employed for rotating the antenna.

The general preference is for horizontal polarization, primarily because less height is required to clear surrounding obstructions when all the antenna elements are in the horizontal plane. This is important at 14 and 28 Mc. where the elements are fairly long.

An easily-constructed supporting frame for a horizontal array is shown in Fig. 10-96. It may be made of 1 × 2-inch lumber, preferably oak, for the center sections *B*, *C*, and *D*. The outer arms, *E*, and cross braces, *F*, may be of white pine or cypress. The square block, *A*, at the center supports the whole structure and may be coupled to the pole by any convenient means which permits rotation. Alternatively, the block may be firmly fastened to the pole and the latter rotated in bearings affixed to the side of the house.

Another type of construction is shown in Fig. 10-97, with details in Figs. 10-98 and 10-99. This method, suitable for 28-Mc. beams, uses a section of ordinary ladder as the main support, with crosspieces to hold the tubing antenna elements. Fig. 10-98 also indicates a method of adjusting the lengths of the parasitic elements and bringing the transmission line down through the supporting pole from a delta match. The latter is especially adapted to construction in which the pole rather than the framework alone is rotated.

Metal Booms

Metal can be used to support the elements of the rotary beam. For 28 Mc., a piece of 2-inch diameter duraluminum tubing makes a good "boom" for supporting the elements. The elements can be made to slide through suitable holes in the boom, or special clamps and brackets can be fashioned to support the elements. The antenna of Fig. 10-84D shows one example of such construction.

Generally it is not practicable to support the elements of a 14-Mc. beam by a single-piece boom, because the size of the elements requires a stronger structure. However, by making use of tubing or duraluminum angle, a lightweight support for a 20-meter antenna can be built. The four-element beam shown in

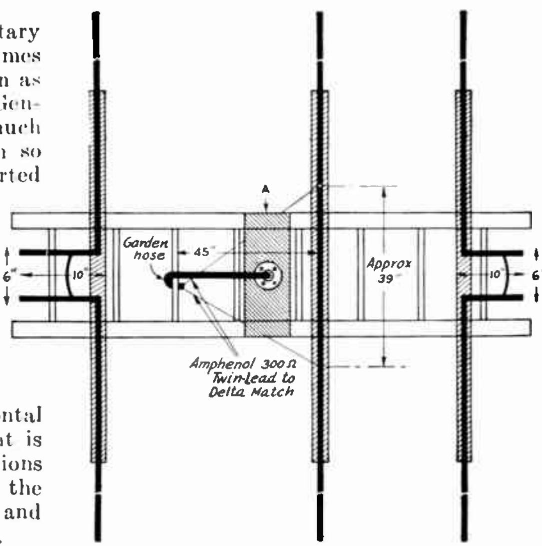


Fig. 10-98 — **Top-view drawing of the ladder support and mounted elements. Lengths of director and reflector are adjusted by means of the shorting bars on the small stubs at the center. The drawing also shows a method for pulling off the wires of a delta match and feeding 300-ohm Twin-Lead transmission line through the pipe support.**

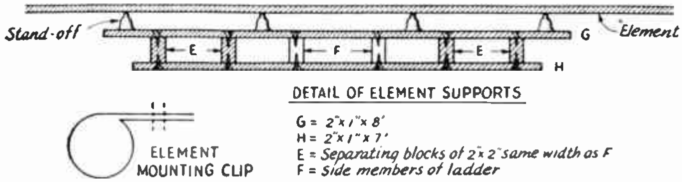


Fig. 10-99 — Detail of element supports for the ladder beam.

Figs. 10-100, 10-101 and 10-102 is an example. It uses 1¾-inch angle for the main pieces and ¾-inch angle for the other members, and the entire framework plus elements weighs only forty pounds. This simplifies considerably the problem of supporting the beam.

The following aluminum pieces are required:

4 — 1-inch diameter tubing, 12 feet long, $\frac{1}{16}$-inch wall
8 — ⅞-inch diameter tubing, 12 feet long, $\frac{1}{32}$-inch wall. Must fit snugly into 1-inch tubing.
2 — 1¾-inch angle, 21 feet long
2 — ¾-inch angle, 21 feet long
4 — ¾-inch angle, 1 foot long
2 — ½-inch diameter tubing, 6 feet long

Aluminum tubing and angle corresponding to the above sizes can possibly be bought from scrap dealers at reasonable prices, if not directly from the manufacturer. If the sections of the elements do not fit snugly, insert shims or make some other provision for a tight fit, since the appearance of the beam will be spoiled by sagging elements. Some amateurs reinforce their beam elements with copper-clad steel wire supported a foot above the elements at the boom and tied to the extreme ends of the elements.

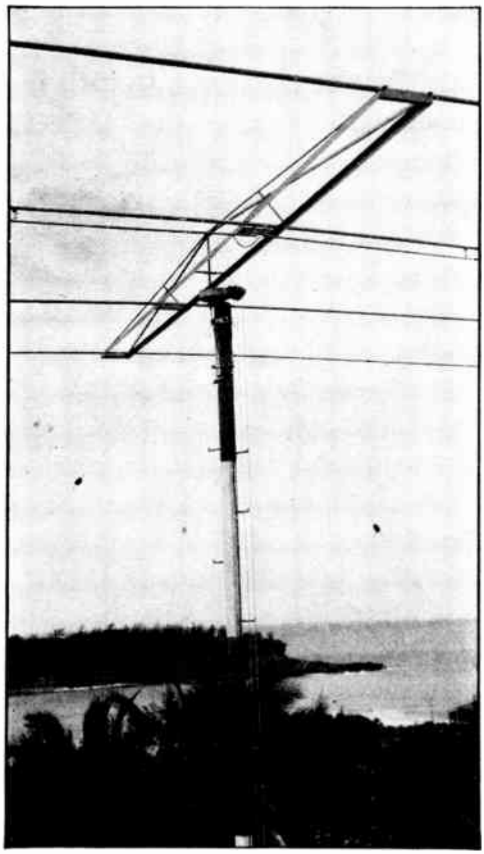


Fig. 10-100 — A four-element 14-Mc. beam of light-weight all-metal construction. Fed by coaxial cable and hand-rotated, the antenna and boom assembly weighs only 40 pounds. (KH6LJ, Dec., 1947, *QST*.)

As shown in Fig. 10-101A, two 1¾-inch aluminum angles 21 feet long serve as the main members of the boom. They are spaced one foot apart. The elements are spaced 7 feet apart. Wooden spacers of 2 × 2 are placed at the end of the boom and screwed on with brass screws. These spacers are also placed under each element where it crosses the boom. These spacers may be unnecessary if the elements are bolted to the boom, but if the construction is as in Fig. 10-101B the spacers are recommended.

The cross braces shown in Fig. 10-102 are put into position at the very last, after the beam is hung in position on the central pivot, since they offer a means for truing up minor sag in the elements.

The central pivot consists of a structure made from ¾-inch angle iron and ½-inch pipe, as shown in Fig. 10-101C. It has to be brazed. The crossbar rest is made separate from the boom and central pivot, and affords a means for tilting the beam when unbolted from these structures. The ½-inch pipe is drilled for the coaxial line that is fed through this pipe. The pinion gear on the ½-inch pipe should be brazed on.

A washing-machine gear train is well suited for this type of beam. Another possibility (used in this instance) is a discarded forge blower. It was fitted with a ½-inch pipe which serves as the central pivot. The gear train ends up in a "V"-pulley, and the beam is easily rotated by a system of ropes and pulleys that ends up in an automobile steering wheel at the operating position. A plumb bob attached to the shaft of the steering wheel serves as a direction indicator. A small cardboard scale mounted along the line of plumb-bob travel can be readily calibrated to show the direction of the beam.

The supporting structure for this beam consists of a 4 × 4 pole 30 feet long, with ten-foot extensions of 2 × 4 bolted to both sides of the bottom, making the total length about 36 feet. Two sets of guy wires should be used, approximately 2 feet and 15 feet from the top. As an alternative, the pole can be set against the side of the house, and only the top set of guys used to provide additional support.

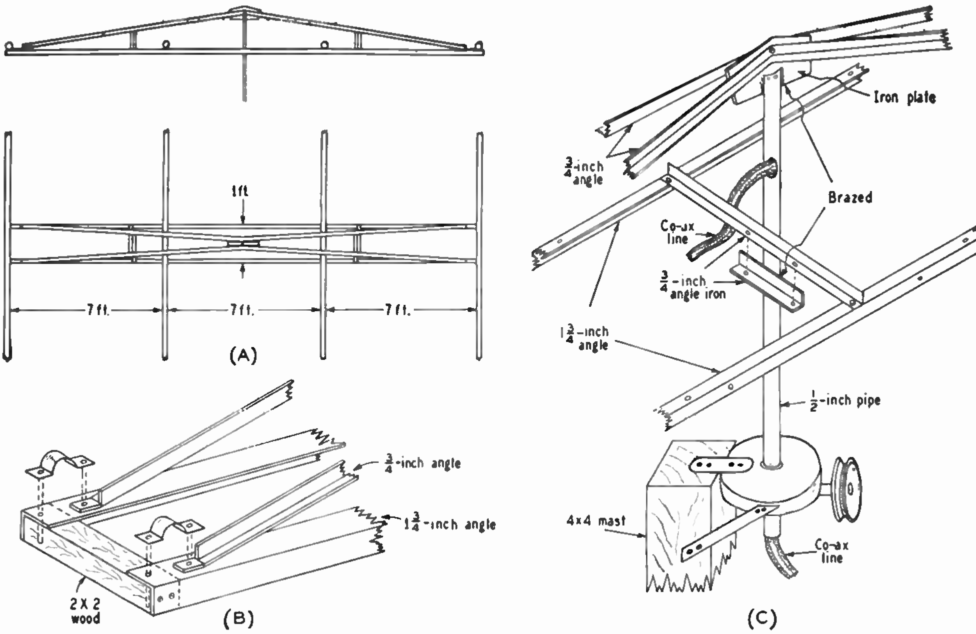


Fig. 10-101 — Details of the 4-element beam construction. The general dimensions and arrangement of the beam are given in A, the detail of the ends of the boom is shown at B, and C shows the construction of the central pivot. A discarded-forge blower gear train is used to drive the assembly.

With all-metal construction, delta match or "T"-match is the only practical matching method to use to the line, since anything else requires opening the driven element at the center, and this complicates the support problem for that element.

A Wooden Boom for 14 Mc.

Many amateurs prefer to build their beam booms from standard pieces of lumber, and the beam shown in Figs. 10-103 and 10-104 is an example of excellent design in wooden-boom construction. The boom members are two 20-foot 2 × 4s fastened to the 4 × 12 × 24-inch center block with six lag screws. The two center screws serve as the axis for tilting — the other four lock the boom in position after final assembly and adjustment have been completed. The blocks midway from each end are 2 × 4s spaced about six inches apart, with a long bolt between them. When this bolt is drawn tight, a very sturdy box brace is formed.

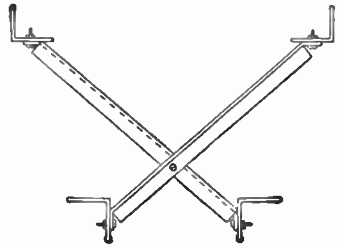


Fig. 10-102 — The boom for the 4-element beam is cross-braced at two points, about 6½ feet in from the ends.

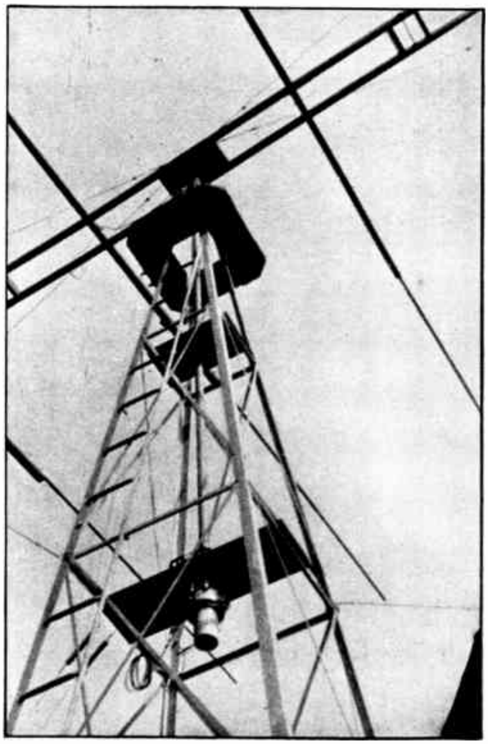


Fig. 10-103 — A wooden boom for a 4-element 14-Mc. boom can be made quite strong by judicious use of guy wires. This installation is made on a windmill tower, and the drive motor is mounted halfway down on the tower. (W6MJB, Nov., 1947, *QST*.)

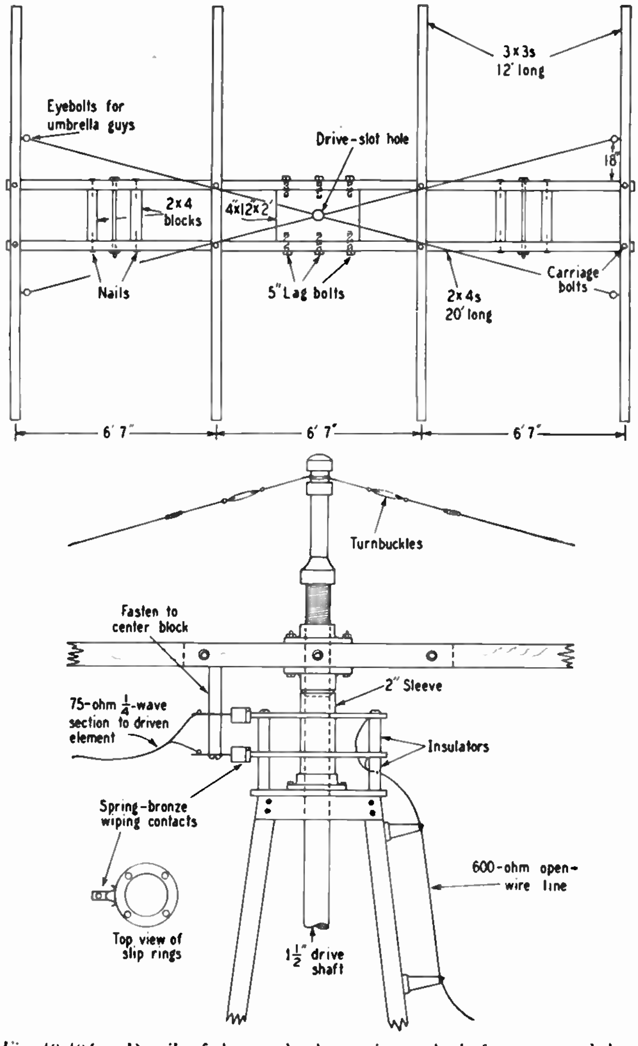


Fig. 10-104 Details of the wooden boom, its method of support and the construction of the slip rings.

The crossarms are 3 × 3s twelve feet long, bolted to the boom with carriage bolts.

The umbrella guys should have turnbuckles in them, and the guys are fastened to the center support after the beam has been permanently locked in its horizontal position. With the turnbuckles properly adjusted, there will be no sag in the boom, the elements will be parallel and neat, and weaving in the wind will be eliminated.

The elements are 1⅜- and 1½-inch diameter duralumin tubing, supported by 1½-inch standoff insulators. Hose clamps are used to hold the elements on the insulators. Final adjustment of element lengths is possible through "hairpin" loops. The tower for the beam shown in Fig. 10-103 was a Sears-Roebuck windmill tower. The driving motor for the beam was located halfway down the tower, the torque being transmitted through a length of 1½-inch drive shaft. A pipe flange is welded to the drive shaft and bolted to the center block. A cone bearing is obtained by turning both the flange and a sleeve of 2-inch pipe to match, as shown in Fig. 10-104.

One method of matching the line to the antenna is to use a quarter wavelength of 75-ohm Twin-Lead between the radiator and the slip-ring contacts, to match a 600-ohm line from the slip rings to the transmitter.

A 600-ohm open-wire line is run to a point about halfway up on the tower, then up the side of the tower to the slip rings. The slip rings are mounted on the top of the tower, directly under the center block. A quarter-wave-length matching section of transmitting-type 75-ohm Amphenol Twin-Lead hangs in a loop between the driven element and the slip-ring contacts.

"Plumber's-Delight" Construction

The lightest beam to build is the so-called "plumber's delight" — an array constructed entirely of metal, with no insulating members between the elements and the supporting structure. Suggested constructional details are shown in Figs. 10-105, 10-106, 10-107, 10-108 and 10-109.

The boom can be built of two lengths of 3-inch diameter 24ST dural tubing of 0.072-inch wall thickness, as shown in Fig. 10-105. The two sections are spliced together with a three-foot length of 6 × 6 oak, turned down at each end to fit inside the tubing. The center of the block is left square to provide a flat surface to attach to the vertical rotating pipe. At each extremity of this boom is cut a hole the exact diameter of the parasitic elements. A two-foot length of ¾-inch pipe, complete with flange mounting plate,

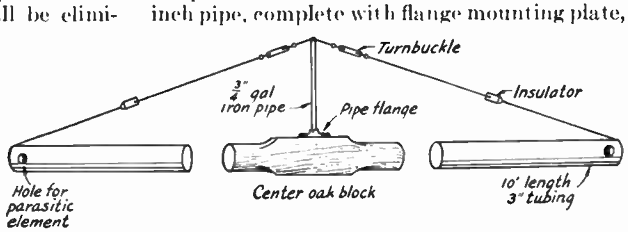


Fig. 10-105 — The boom is made of two 10-foot lengths of dural tubing slipped over a 3-foot oak block and held in place with 2-inch wood screws. Guy wires from the center add strength to the boom structure.

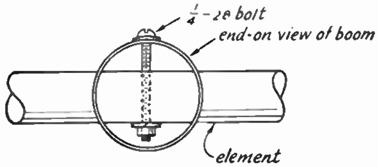


Fig. 10-106 — The center element section is held in the boom with a ¼-28 machine screw, nut and lock washer. The guy wire attaches to the head of the bolt.

is bolted to the top surface of the oak block, and a single guy wire is run to each end of the boom. An egg insulator and a turnbuckle are placed in each guy. The turnbuckles should be tightened until there is no sag in the boom when it is supported at the center, and then safety-wired. Finally the center block should be given a good coat of paint or varnish.

The elements can be made of three 12-foot lengths of dural tubing, the two outside lengths telescoping inside the center section. The ends of the center section should be slotted for a distance of about 4 inches with a hack saw, but it is advisable to do the slotting after the center sections have been assembled on the boom. The parasitic-element center sections are fastened to the boom with ¼-inch bolts, as shown in Fig. 10-106, while the driven element is secured in a cradle made of half sections of iron pipe welded together, as shown in Fig. 10-107. The cradle is bolted to the boom with three ¼-inch bolts, and the driven element is held fast with two bolts or with adjustable aircraft-tubing clamps.

The feedline for the antenna can be any balanced line, of from 200 to 600 ohms imped-

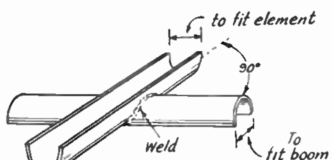


Fig. 10-107 — The clamp for the driven element is made by splitting 1-foot lengths of iron pipe and welding them as shown.

ance, and it is most conveniently coupled through a "T"-match. This "T"-match assembly can be made from two 4-foot lengths of dural tubing joined together by a piece of broomstick, as shown in Fig. 10-109. The "T" is connected to the antenna by two clamps fashioned of 1-inch-wide brass strip.

A convenient method for supporting the boom atop the pipe used to rotate the beam is shown in Fig. 10-108. A "U"-channel into which the boom will fit is welded to the end of the pipe. Holes are drilled in the side of the channel corresponding to holes in the boom. The boom is hoisted up and positioned between the two flanges and a bolt run through the flanges and the boom. The boom can then be swung into a horizontal position and the second bolt put in place.

Feeder Connections

For beams that rotate only 180 degrees, it is relatively simple to bring off feeders by making a short section of the feeder, just where it leaves the rotating member, of flexible wire. Enough slack should be left so that there is no danger of breaking or twisting. Stops should be placed on the rotating shaft of the antenna so that it will be impossible for the feeders to "wind up." This method also can be used with antennas that rotate the full 360 degrees, but again a stop is necessary to avoid jamming the feeders.

For continuous rotation, the sliding contact is simple and, when properly built, quite prac-

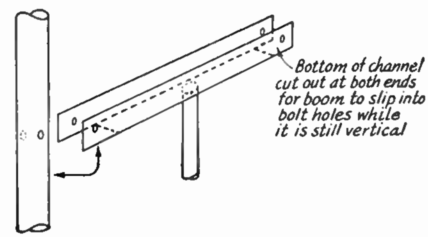


Fig. 10-108 — The mounting plate is made from a length of "U"-channel iron cut and drilled as shown. The boom is raised vertically until one set of bolt holes is in line and a bolt is slipped through. The boom is then swung into its horizontal position and the other bolt is put in place.

ticable. Fig. 10-110 shows two methods of making sliding contacts. The chief points to keep in mind are that the contact surfaces should be wide enough to take care of wobble in the rotating shaft, and that the contact surfaces should be kept clean. Spring contacts are essential, and an "umbrella" or other scheme for keeping rain off the contacts is a desirable addition. Sliding contacts preferably should be used with nonresonant open lines where the characteristic impedance is of the order of 500 to 600 ohms, so that the line current is low.

The possibility of poor connections in sliding contacts can be avoided by using inductive coupling at the antenna, with one coil rotating on the antenna and the other fixed in position, the two coils being arranged so that the coupling does not change when the antenna is rotated. Such an arrangement is shown in Fig. 10-111, adapted to an antenna system in which the pole

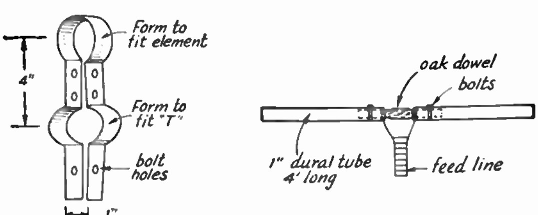


Fig. 10-109 — Details of the "T"-match assembly.

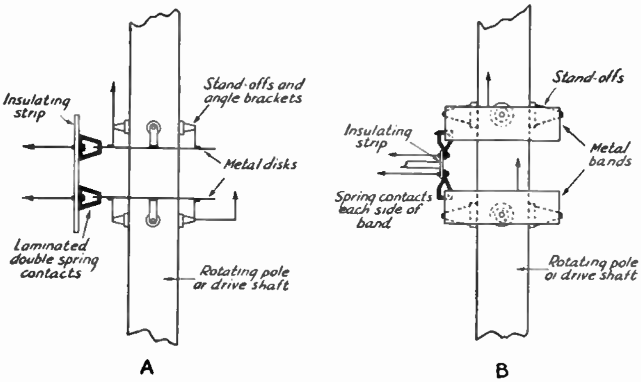


Fig. 10-110 — **Ideas in sliding contacts for rotatable antenna feeder connection to permit continuous rotation. The broad bearing surfaces take care of any wobble in the rotating mast or driving shaft.**

itself rotates. A quarter-wave feeder system is connected to a tuned pick-up circuit whose inductance is coupled to a link. In the drawing, the link coil connects to a twisted-pair transmission line, but any type of line such as flexible coaxial cable can be used. The circuit would be adjusted in the same way as any link-coupled circuit, and the number of turns in the link should be varied to give proper loading on the transmitter. The rotating coupling circuit of course tunes to the transmitting frequency. The whole thing is equivalent to a link-coupled antenna tuner mounted on the pole, using a parallel-tuned tank at the end of a quarter-wave line to center-feed the antenna. To maintain constant coupling, the two coils should be quite rigid and the pole should rotate without wobble. The two coils might be made a part of the upper bearing assembly holding the rotating pole in position.

Other variations of the inductive-coupled system can be worked out. The tuned circuit might, for instance, be placed at the end of a 600-ohm line, and a one-turn link used to couple directly to the center of the antenna, if the construction of the rotary member permits. In this case the coupling can be varied by changing the L/C ratio in the tuned circuit. For mechanical strength the coupling coils preferably should be made of ¼-inch copper tubing, well braced with insulating strips to keep them rigid.

Rotation

It is convenient to use a motor to rotate the beam, but it is not always necessary, especially if a rope-and-pulley arrangement can be brought into the operating room. If the pole can be mounted near a window in the operating room, hand rotation of the beam will work out quite well, as has been proven by many amateur installations.

If the use of a rope and pulleys is impracticable, motor drive is about the only alternative. There are several complete motor-driven rotators on the market, and they are easy to mount, convenient to use, and require little or no maintenance. However, to many the cost of such units puts them out of reach, and a homemade unit must be considered. Generally speaking, lightweight units are better because they reduce the load on the mast or tower.

The speed of rotation should not be too great — one or two r.p.m. is about right. This requires a considerable gear reduction from the usual 1750-r.p.m. speed of small induction motors; a large reduction is advantageous because the gear train will prevent the beam from turning in weather-vane fashion in a wind. The ordinary structure does not require a great deal of power for rotation at slow speed, and a ⅛-hp. motor will be ample. Even small series motors of the sewing-machine type will develop enough power to turn a 28-Mc. beam at slow speed. If possible, a reversible motor should be used so that it will not be necessary to go through nearly 360 degrees to bring the beam back to a direction only slightly different, but in the opposite direction of rotation, to the direction to which it may be pointed at the moment. In cases where the pole is stationary and only the supporting framework rotates, it will be necessary to mount the motor and gear train in a housing on or near the top of the pole. If the pole rotates, the motor can be installed in a

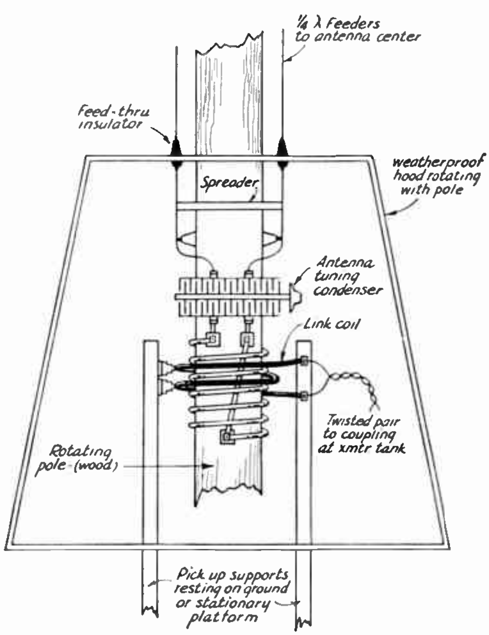


Fig. 10-111 — **One method of transmission line-antenna system coupling which eliminates sliding contacts. The low-impedance line is link-coupled to a tuned line.**

more accessible location (see Fig. 10-103).

Parts from junked automobiles often provide gear trains and bearings for rotating the antenna. Rear axles, in particular, can readily be adapted to the purpose. Driving motors and gear housings will stand the weather better if given a coat of aluminum paint followed by two coats of enamel and a coat of glyptal varnish. Even commercial units will last longer if treated with glyptal varnish. Be sure, of course, that the surfaces are clean and free from grease before painting them. Grease can be removed by brushing it with kerosene and then squirting the surface with a solid stream of water. The work can then be wiped dry with a rag.

If hand rotation of the beam is used, or if the rotating motor drives the beam through a pulley system, bronze cable or chain drive is preferable to rope. However, if you must use rope, be sure to soak it overnight in pure linseed oil and then let it dry for several days before permanent installation.

The power and control leads to the rotator should be run in electrical conduit or in lead covering, and the metal should be grounded. Often r.f. appearing in power leads can be reduced by suitable filtering, but running wires in conduit is generally easier and more satisfactory. Any r.f. in the wiring can sometimes be responsible for feed-back in a 'phone transmitter. "Hash" from the motor is also reduced by shielding the wires, but it is often necessary to install a small filter at the motor to reduce this source of interference. Motor noise appearing in the receiver is a nuisance, since it is usual practice to determine the proper direction for the beam by rotating it while listening to the station it is desired to work and setting the antenna at the point that gives maximum signal strength.

The outside electrical connections should be soldered, bound with rubber tape followed by regular friction tape, and then given a coat of glyptal varnish.

About V.H.F.

In the days when DX activity first burgeoned on our lower frequencies the assignments above 30 Mc. were not too highly regarded. It was assumed that propagation on these frequencies was limited to distances only slightly beyond the visual horizon, and thus the bands allocated to amateurs in this region were used principally in areas where large concentrations of population brought hundreds of workers within local range of one another. In the early thirties activity boomed on 56 Mc. in the larger cities of the United States, but there were few stations elsewhere. Use of frequencies higher than 60 Mc. was confined to a few experimentally-inclined amateurs here and there.

In 1934, '35 and '36, new types of propagation were discovered by amateurs, and the opportunities for v.h.f. DX so brought to light caused a tremendous growth in activity, particularly in areas where it had not previously existed. Up to this time, practically all v.h.f. work had been done with the simplest sort of gear, mainly modulated-oscillator transmitters and superregenerative receivers; but when our available space began to fill with DX signals it became obvious that, if we were to realize anything like the possibilities inherent in this type of work, we must have improved techniques, whereby more stations could be accommodated in a given area. Crystal-controlled transmitters and superheterodyne receivers, permitting utilization of the 56-Mc. band on a scale comparable with that obtaining on lower frequencies, became the order of the day, and by the end of 1938 stabilization of transmitters used on all frequencies up to 60 Mc. became mandatory.

With the impetus of improved techniques, operating ranges on 56 Mc. grew by leaps and bounds. Meanwhile the use of the simplest form of equipment was transferred to the next higher band, then 112 Mc.; and this band, in turn, took over the burden of heavy urban occupancy formerly carried by the 5-meter band. Soon our principal cities were teeming with 112-Mc. activity, and before long it was found that this band, too, had much of interest to offer. Even more than had been the case on 56 Mc., it was found that weather conditions had a profound effect on 112-Mc. propagation, and before the close-down of amateur activity, at the entry of our country into the war, the record for 112-Mc. work had passed the 300-mile mark. There was a smattering of activity on the still higher frequencies of 224 and 400 Mc. as well.

In the postwar years the value of the very-high frequencies has been amply demonstrated. World-wide communication has been accomplished on 50 Mc.; two-way work on 144 Mc. has been extended to more than 800 miles; and pioneering effort on 220 and 420 Mc. is establishing these bands as fields of great interest for the experimentally-inclined amateur. The v.h.f. worker need no longer apologize for his interests. His frequencies are among the most highly prized in the entire spectrum, and his is now regarded as one of the major fields of amateur endeavor.

Propagation Phenomena

A thorough understanding of the basic principles of wave propagation, outlined in Chapter Four, is a most useful tool for the v.h.f. worker. Much of the pleasure and satisfaction to be derived from v.h.f. endeavor lies in making the best possible use of propagation vagaries resulting from natural phenomena. Contrary to the impression of many newcomers to the field, a working knowledge of v.h.f. propagation is not difficult of attainment. Below are listed the principal ways by which v.h.f. waves may be propagated over abnormal distances.

F_2-Layer Reflection

The "normal" contacts made on 28 Mc. and lower frequencies are the result of reflection of the transmitted wave by the F_2 layer, the ionization density of which varies with solar activity, the highest frequencies being reflected at the peak of the 11-year solar cycle. The maximum usable frequency (m.u.f.) for F_2 reflection also rises and falls with other well-defined cycles, including daily, monthly, and seasonal variations, all related to conditions on the sun and its position with respect to the earth.

At the low point of the 11-year cycle, such as the period we were entering at the outbreak of war, the m.u.f. may reach 28 Mc. only during a short period each spring and fall, whereas it may go to 60 Mc. or higher at the peak of the cycle. The fall of 1946 saw the first authentic

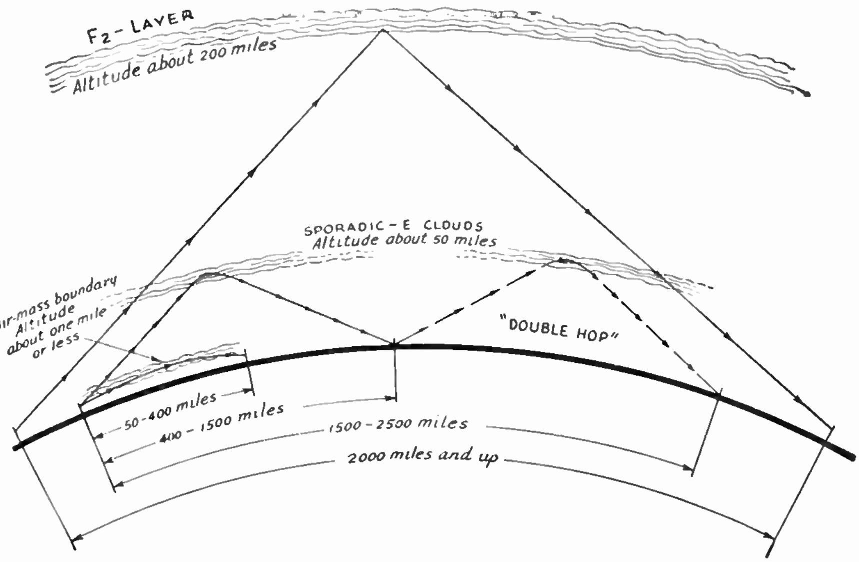


Fig. 11-1 — The principal means by which v.h.f. waves may be returned to earth. The F_2 layer, highest of the known ionospheric layers, is capable of reflecting 50-Mc. signals during the period around the peak of the 11-year solar cycle, and may support communication over world-wide distances. Sporadic ionization of the E layer produces "short-skip" contacts at medium distances. It is a fairly frequent occurrence regardless of the solar cycle, but is most common in May through August. Refraction of v.h.f. waves also takes place at air-mass boundaries in the lower atmosphere, making possible reception of signals at distances up to 300 miles or more without a skip zone.

instances of long-distance 50-Mc. work by this medium, and it is probable that F_2 DX will be workable on 50 Mc. until about 1950. In the northern latitudes there are peaks of m.u.f. each spring and fall, with a low period during the summer and a slight dropping-off during the midwinter months. At or near the Equator conditions are more or less constant at all seasons.

Fortunately the F_2 m.u.f. is quite readily determined by observation, and means are available whereby it may be estimated quite accurately for any path at any time. It is predictable for months in advance,[1] enabling the v.h.f. worker to arrange test schedules with distant stations at propitious times. As there are numerous signals, both harmonics and fundamental transmissions, on the air in the range between 28 and 50 Mc., it is possible for an observer to determine the approximate m.u.f. by careful listening in this range. A series of daily observations will serve to show if the m.u.f. is rising or falling from day to day, and once the peak for a given month is determined it can be assumed that the peak for the following month will occur about 27 days later, this cycle coinciding with the turning of the sun on its axis. The working range, via F_2 skip, will be roughly comparable to that on 28 Mc., though the *minimum* distance is somewhat longer. Two-way work on 50 Mc. by means of reflection from the F_2 layer has been accomplished over distances ranging from 2200 to 10,500 miles. The maximum frequency for F_2 reflection is believed to be in the vicinity of 70 Mc.

[1] *Basic Radio Propagation Predictions*, issued monthly, three months in advance, by the Central Radio Propagation Laboratory of the National Bureau of Standards. Order from the Supt. of Documents, Washington 25, D. C.; $1.00 per year.

Sporadic-E Skip

Patchy concentrations of ionization in the E-layer region are often responsible for reflection of signals on 28 and 50 Mc. This is the popular "short skip" that provides fine contacts on both bands in the range between 400 and 1300 miles. It is most common in May, June and July, during the early evening hours, but it may occur at any time or season. Since it is largely unpredictable, at our present state of knowledge, sporadic-E skip is of high "surprise value." Multiple-hop effects may appear, when ionization develops simultaneously over large areas, making possible work over distances of more than 2500 miles. The known limit of sporadic-E skip is about 100 Mc., but rare cases of 144-Mc. reception at 1000 to 1200 miles indicate that E-layer reflection may be possible in that band.

Aurora Effect

Low-frequency communication is occasionally wiped out by absorption of these frequencies in the ionosphere, when ionospheric storms, associated with variations in the earth's magnetic field, occur. During such disturbances, however, 50-Mc. signals may be reflected back to earth, making communication possible over distances not normally workable on this band. Magnetic storms may be accompanied by an aurora-borealis display, if the disturbance occurs at night and visibility is good. When the aurora is confined to the northern sky, aiming a directional array at the auroral curtain will bring in 50-Mc. signals strongest, regardless of the true direction to the transmitting station. When the display is widespread there may be only a slight improvement noted when the array is aimed north. The latter condition is often noticed during the period around the peak of the 11-year cycle, when solar activity is spread well over the sun's surface, instead of being concentrated in the region near the solar equator.

Aurora-reflected signals are characterized by a rapid flutter, which lends a "dribbling" sound to 28-Mc. carriers and may render modulation on 50-Mc. signals completely unreadable. The only satisfactory means of communication then becomes straight c.w. The effect may be noticeable on signals from any distance other than purely local, and stations up to about 500 miles in any direction may be worked at the peak of the disturbance. Unlike the two methods of propagation previously described, aurora effect exhibits no skip zone. It has been observed mainly on frequencies up to about 60 Mc., though there have been several instances when it has shown up on 144 Mc. The highest frequency for aurora reflection is not yet known.

Reflections from Meteor Trails

Probably the least-known means of v.h.f. wave propagation is that resulting from the passage of meteors across the signal path. Reflections from the ionized meteor trails may be noted as a Doppler-effect whistle on the carrier of a signal already being received, or they may cause bursts of reception from stations not normally receivable. Sudden large increases in strength of normally-weak signals are another manifestation of this effect. Ordinarily such reflections are of little value in extending communication ranges, since the increases in signal strength are of short duration, but meteor showers of considerable magnitude and duration may provide fluttery 50-Mc. signals from distances up to 1000 miles or more. Signals so reflected have a combination of the characteristics of aurora and sporadic-*E* skip.

Tropospheric Bending

Refraction of radio waves takes place whenever a change in refractive index is encountered. This may occur at one of the ionized layers of the ionosphere, as mentioned above, or it may exist at the boundary area between two different types of air masses, in the region close to the earth's surface. A warm, moist air mass from over the Gulf of Mexico, for instance, may overrun a cold, dry air mass which may have had its origin in northern Canada. Each tends to retain its original characteristics for considerable periods of time, and there may be a well-defined boundary between the two for as much as several days. When such an air-mass boundary exists near the midpoint between two v.h.f. stations separated by 50 to 300 miles or more, a considerable degree of refraction takes place, and signals run high above the average value. Under ideal conditions there may be almost no attenuation, and signals from far beyond the visual horizon will come through with strength comparable to that of local stations.

Many factors other than air-mass movement of a continental character may provide increased v.h.f. operating range. The convection that takes place along our coastal areas in warm weather is a good example. The rapid cooling of the earth after a hot day in summer, with the air aloft cooling more slowly, is another, producing a rise in signal strength in the period around sundown. The early-morning hours, when the sun heats the air aloft, before the temperature of the earth's surface begins its daily rise, may frequently be the best hours of the day for extended v.h.f. range, particularly in clear, calm weather, when the barometer is high and the humidity low.

Any weather condition that produces a pronounced boundary between air masses of different temperature and humidity character-

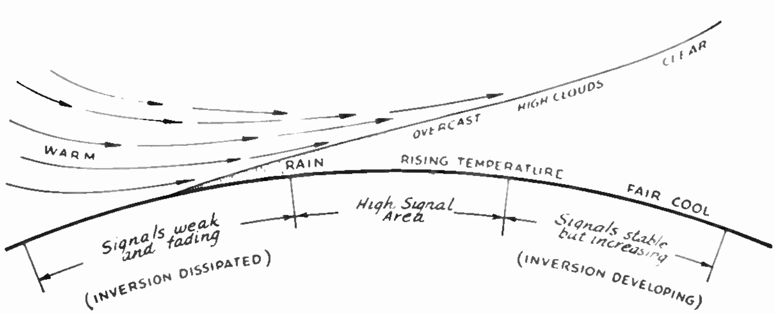


Fig. 11-2 — Illustrating a typical weather sequence, with associated variations in v.h.f. propagation. At the right is a cold air mass (fair weather, high or rising barometer, moderate summer temperatures). Approaching this from the left is a warm moist air mass, which overruns the cold air at the point of contact, creating a temperature inversion and considerable bending of v.h.f. waves. At the left, in the storm area, the inversion is dissipated and signals are weak and subject to fading. Barometer is low or falling at this point.

V.H.F. Receivers

In its basic principles, modern receiving equipment for 50, 144 and even 220 Mc. differs very little from that used on lower amateur frequencies. Federal regulations impose identical restrictions on all frequencies below 54 Mc. as to stability of transmitted signals, and experience has shown that only through the use of stabilized transmitters and selective receivers can the full possibilities of 144 and 220 Mc. be realized. Thus, for AM or NFM reception, at least, receivers for the v.h.f. bands may have the same selectivity as those used for lower frequencies.

This order of selectivity is not only possible but desirable, since it permits a considerable increase in the number of stations that can work in a band without harmful interference. High selectivity also aids greatly in improving the signal-to-noise ratio, both as to noise originating in the receiver itself and its response to external noise. The effective sensitivity of a receiver having "communications" selectivity can be made considerably higher than is possible with nonselective receivers. First on the old 56-Mc. band in the late '30s, then on 144 Mc. in the early part of the postwar period, and currently on 220 Mc., the change to selective superheterodynes for use at the more progressive v.h.f. stations marked the beginning of real extensions of the effective operating radius of v.h.f. stations.

Superregenerative receivers, once the most popular type for v.h.f. work, are now used mainly for portable operation, or for other applications where maximum selectivity and sensitivity are not required. Its lack of these essential features, its inability to provide satisfactory reception of FM signals, and its tendency to radiate a strong interfering signal, rule the superregenerator out as a home-station receiver in areas where there is appreciable activity on the v.h.f. bands.

Superheterodynes for V.H.F.

Superheterodynes for 50 Mc. and higher should have fairly-high intermediate frequencies to reduce both image response and oscillator "pulling." For example, a difference between signal and image frequencies of 900 kc. (the difference when the i.f. is 450 kc.) is a very small percentage of the signal frequency; consequently, the response of the r.f. circuits to the image frequency is nearly as great as to the desired frequency. To obtain discrimination against the image equal to that obtainable at 3.5 Mc. would require an i.f. 16 times as high, or about 7 Mc. However, the Q of tuned circuits is less in the v.h.f. range than it is at lower frequencies, chiefly because the tube loading is considerably greater, and thus still higher intermediate frequencies are desirable. A practical compromise is reached at about 10 Mc., and the standard i.f. for converters and commercial v.h.f. receivers is 10.7 Mc.

To obtain the desired degree of selectivity with a reasonable number of i.f. stages, the double-conversion principle is often employed. A 10-Mc. intermediate frequency, for example, is changed to an i.f. of 1600 or 455 kc. by adding a second mixer-oscillator combination.

Most v.h.f. receivers are of this category, general practice being to use a conventional communications receiver to handle the i.f. output of a relatively-simple converter. Even a broadcast receiver which has a "short-wave band" may be used as an i.f. amplifier in this manner with good results. Only crystal-controlled or otherwise stabilized signals can be received with such combinations, but since nearly all v.h.f. amateur stations now employ stabilized transmitters this is not likely to be troublesome.

When a high-selectivity i.f. is employed in v.h.f. reception, the stability of the oscillator is a primary problem, and care must be taken to be sure that the converter oscillator is both mechanically and electrically stable. One satisfactory solution to this problem is the use of a crystal-controlled oscillator and frequency multiplier to supply the injection voltage, the method used in the 144-Mc. converter shown in Figs. 12-10–12-12.

Where reception of wide-band FM or unstable signals of the modulated-oscillator type is desired, a converter may be used ahead of an i.f. of the type used for FM broadcast reception, or with a complete receiver of the FM broadcast variety. A superregenerative detector operating at the intermediate frequency, with or without additional i.f. amplifier stages, also may serve as an i.f. and detector system for reception of wide-band signals. By using a high i.f. (10 to 30 Mc. or so) and by resistive loading of the i.f. transformers, almost any desired degree of bandwidth can be secured, providing good voice quality on all but the most unstable signals. Any of these methods may be used for reception in the u.h.f. and microwave regions, where stabilized transmission is extremely difficult at the current state of the art.

A Two-Tube Converter for 50 Mc.

The converter shown in Figs. 12-1–12-4 is designed to provide good performance on 50 Mc. with a minimum of complication. It employs a 6AK5 tuned r.f. stage and a dual-triode mixer-oscillator using a 12AT7. It has its own built-in power supply. To reduce tracking problems and simplify construction, only the oscillator is tuned by means of the vernier dial. The r.f. amplifier grid circuit and the mixer grid circuit, neither of which is critical in its tuning, are ganged together, and are adjusted by means of the knob at the left of the vernier dial, as seen in Fig. 12-1. In actual operation this control may be peaked at about 51 Mc. and the converter tuned over the lower half of the band or more before any readjustment of the knob is required.

Electrical and Mechanical Details

One section of the 12AT7 is used as a triode oscillator, employing a Colpitts circuit. It covers a range of six megacycles, in order to permit tuning well below the low end of the 50-Mc. band, a useful feature when one is interested in searching these frequencies for signs of DX. By resetting C_4 and C_5 to a higher capacitance the tuning range may be extended down to about 45 Mc., at the sacrifice of the top three megacycles of the 6-meter band.

The oscillator is tuned with a standard split-stator variable capacitor, C_3, the rotor of which is grounded. Parallel capacitance, for increased stability, is supplied by the small air padders, C_4, C_5. The oscillator tank coil, L_5, is attached directly to the stator terminals of the tuning condenser, and its center-tap provides three-point suspension, resistor R_8 being attached to a feed-through bushing directly below it. To prevent microphonics resulting from vibration, the turns of the coil are cemented together at four points.

The 6AK5 and 12AT7 tubes are mounted in an inverted position, with their sockets above the chassis, in order to provide the shortest possible leads. The socket for the 6AK5 has the baffle plate between the two tuned circuits passing across the middle of the socket in such a way that Pins 1 through 4 are on the enclosed side, while the others, which are concerned with the output circuits, are on the side toward the panel. The shield is made of sheet copper and is soldered to the cylindrical shield at the center of the 6AK5 socket. A high value of L/C ratio is maintained in the r.f. and mixer stages by the elimination of the parallel padder condensers which would have been required for tracking, if the oscillator had been tuned from the same shaft.

The intermediate frequency may be any convenient value, and the i.f. transformer is made plug-in so that the frequency may be changed over a wide range if desired. With the values of inductance and capacitance given the i.f. can be varied from about 7 to 11 Mc. Originally 10.7 Mc., the RMA standard i.f. for converters, was tried, but some image trouble was experienced from strong local 10-meter stations, so the i.f. transformer was retuned to 7.4 Mc. Note the position of C_6, the mica trimmer visible in the top-view photograph. It was found necessary to mount this trimmer directly on the 12AT7 terminals, as the mixer oscillated when the trimmer was connected across the coil terminals below the chassis.

Some precautions may also be required to prevent oscillation in the r.f. stage. Note the manner in which the screen and cathode circuits of the 6AK5 are by-passed. Referring to the schematic diagram, Fig. 12-2, it may be seen that both cathode terminals of the 6AK5 are by-passed, the bias resistor, R_1, and its by-pass, C_7, being on the input side, while C_8 is on the opposite side of the shield. Originally all the cathode connections were made to Pin 7,

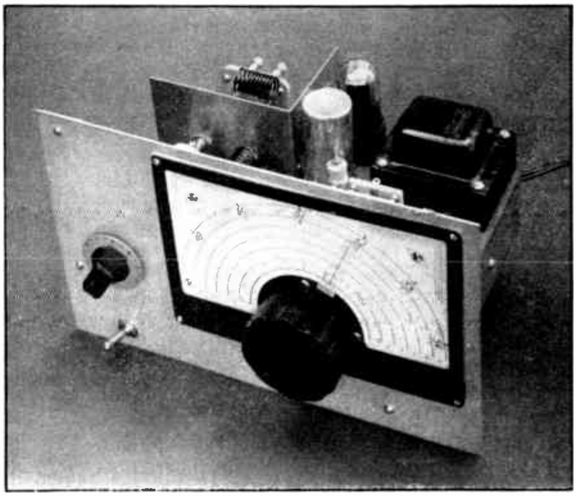


Fig. 12-1 — A two-tube converter for 50 Mc. Only the oscillator is tuned by the vernier dial, simplifying tracking problems. Mixer and r.f. circuits are adjusted by the knob at the left. The unit has a self-contained regulated power supply.

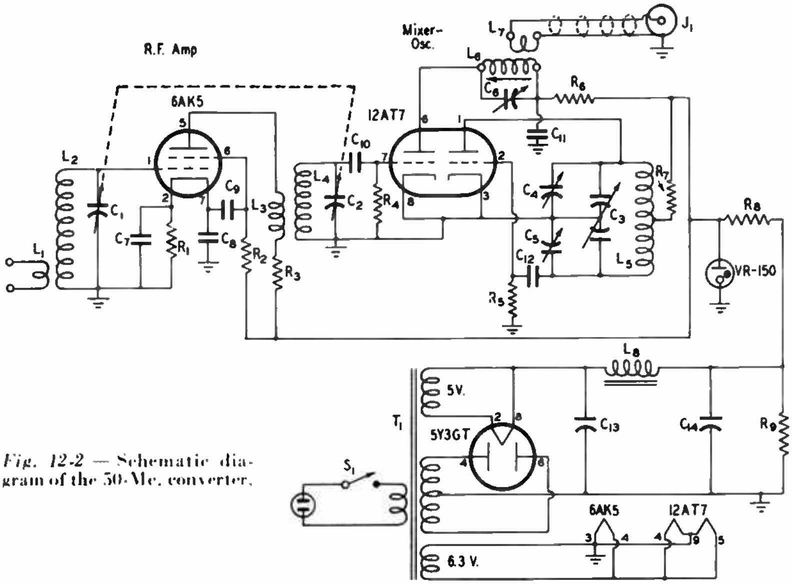


Fig. 12-2 — Schematic diagram of the 50-Mc. converter.

C_1, C_2 — 15-μμfd. variable (Hammarlund MC-20-S with one stator plate removed).
C_3 — 15-μμfd.-per-section split stator (Cardwell ER-15-AD).
C_4, C_5 — 3–30-μμfd. air trimmer (Silver type 619).
C_6 — 3–30-μμfd. mica trimmer (See text for mounting position).
C_7, C_8, C_9 — 680-μμfd. mica.
C_{10}, C_{12} — 100-μμfd. ceramic.
C_{11} — 0.001-μfd. mica.
C_{13}, C_{14} — 10-μfd. 450-volt dual electrolytic.
R_1 — 150 ohms, ½ watt.
R_2, R_5, R_6, R_7 — 10,000 ohms, ½ watt.
R_3 — 2200 ohms.
R_4 — 1.0 megohm.
R_8 — 5000 ohms, 10 watts.
R_9 — 25,000 ohms, 10 watts.

L_1 — 4 turns No. 22 enamel, interwound in cold end of L_2.
L_2 — 10 turns No. 16 enamel, ½-inch inside diam., 1⅛ inches long.
L_3 — 5 turns No. 22 enamel, interwound in cold end of L_4.
L_4 — 9 turns No. 16 enamel, ½-inch inside diam., ⅞ inch long.
L_5 — 11 turns No. 12 enamel, center-tapped, 1⅜ inches long. Cement turns together — see text.
L_6 — 28 turns No. 24 d.s.c., close-wound.
L_7 — 4 turns No. 24 d.s.c. close-wound over cold end of L_6 with one layer of insulating tape between windings.
L_8 — Midget filter choke.
J_1 — Coaxial fitting (Jones S-201).
S_1 — S.p.s.t. toggle switch.
T_1 — Midget power transformer (Thordarson T-13R19).

and the screen was by-passed to ground. With this arrangement the r.f. stage oscillated, so the capacitors were rearranged so that the screen was by-passed (by C_9) to the cathode, Pin 7, and both cathode pins by-passed separately. Even this way the stage may oscillate when no antenna is connected, but in normal operation it is completely stable.

Adjustments

Because of the separate tuning of the oscillator stage, alignment of the converter presents no great problem. First the oscillator should be set to cover the desired range, in this case 40.6 to 46.6 Mc., for a tuning range of 48 to 54 Mc. with an i.f. of 7.4 Mc. The i.f. transformer may then be set to the proper point, by moving the slug or adjusting C_6. A signal generator is convenient, but the adjustment can be made by listening for a noise peak, if no generator is available.

Next the flexible coupling between C_1 and C_2 should be disengaged so that the r.f. and mixer circuits can be adjusted separately. First, an antenna or signal source should be coupled to the mixer grid coil, L_4, and C_2 peaked for

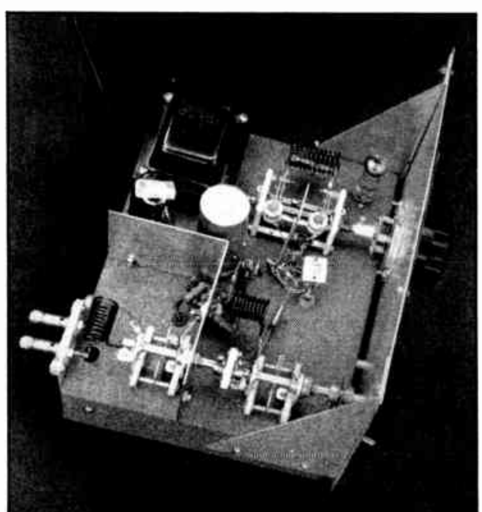


Fig. 12-3 — Top view of the 50-Mc. converter, showing placement of the principal r.f. components. The shielded plug-in coil near the middle of the chassis is the i.f. output transformer. The tube at the upper right of the photo is the voltage regulator.

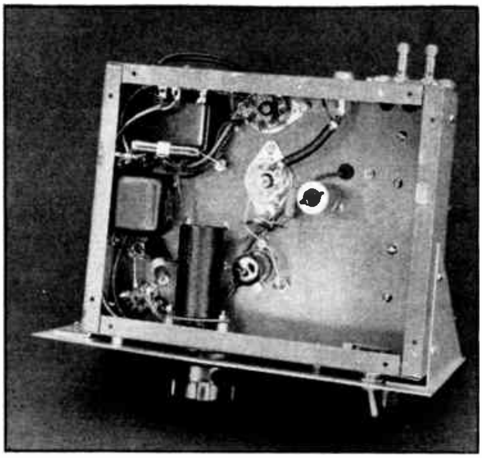


Fig. 12-4 — Bottom view of the 50-Mc. converter, showing the power-supply components. Note that the r.f. and mixer tubes are mounted in an inverted position below the chassis, permitting short leads to the r.f. components above.

maximum signal (or noise) near the middle of the band. Then connect the signal source to the antenna terminals and adjust C_1 in the same manner. If the two grid coils are the proper size the settings of C_1 and C_2 should come out the same. If they do not, the spacing of the turns in L_2 should be adjusted so that the setting of C_1 matches that of C_2.

If the physical arrangement of the converter components is different from that shown in the photographs, it may be necessary to add a small amount of oscillator injection for best performance. This may be done by connecting a short piece of insulated wire to the plate terminal of the oscillator, Pin 1, and running it over to the grid terminal of the mixer, Pin 7. Bend the wire near to the mixer terminal, and adjust its position to give the desired degree of injection. The wire may then be fastened in place with a drop of household cement. Oscillator injection may also be adjusted by using more capacity coupling than is needed, and then increasing the value of the oscillator plate dropping resistor until the desired performance is attained. The optimum degree of coupling is the largest value that can be used without resulting in a change in oscillator frequency when the mixer circuit is tuned.

Reducing Spurious Responses

In locations where there is broadcasting in the high FM band, 50-Mc. receivers and converters may experience severe interference resulting from these high-band signals beating with the second harmonic of the converter oscillator. The selectivity of the r.f. circuits is not sufficiently high at these frequencies to eliminate the unwanted signals, but the interference may be reduced by other means.

First, the output of the converter oscillator should be held to the minimum required to give satisfactory injection. In the case of the converter described above this may be accomplished by making the value of R_7 as high as possible, while still retaining satisfactory performance. This can best be checked by changing the resistor while listening to a very weak signal.

Second, if the above method does not cure the interference, a 100-Mc. trap may be inserted in series with the antenna pick-up coil, L_1. The trap may be made with an adjustable trimmer, or it may use a small fixed capacitor, in which case adjustment is accomplished by spreading or squeezing the turns.

If several interfering signals are present the trap should be adjusted on the strongest. Tune the signal in with the trap out of the circuit, then insert and tune the trap for maximum rejection of the interference. The sharpness of tuning of the trap (and its ability to reject interference) will depend on the L/C ratio. If only one strong signal must be eliminated, a high-C circuit should be used. If there are several, distributed over a considerable frequency range, more inductance and lower capacitance will provide a broader trap response, at the expense of some rejection at any one frequency.

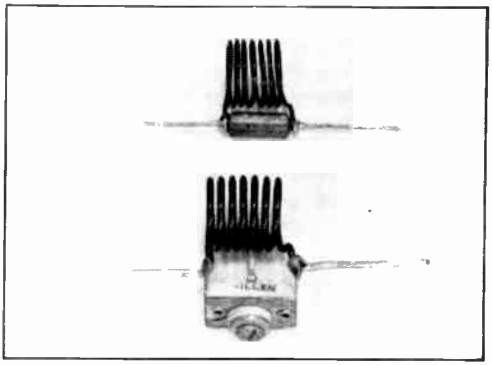


Fig. 12-5 — Examples of 100-Mc. traps used to reduce interference from high-band FM signals.

A usable average value is a coil of 7 turns of No. 14 to 18 wire, ½-inch diameter, spaced about the diameter of the wire. This may be mounted on a mica trimmer (3-30 $\mu\mu$fd.) or on a small 5-$\mu\mu$fd. ceramic condenser.

A Low-Noise Converter for 144 Mc.

The 2-meter converter shown in Figs. 12-6 to 12-9 was designed for superior weak-signal performance, yet it is relatively simple and inexpensive to build. Its r.f. section has a low noise figure, and special attention has been paid to oscillator design, for smooth tuning and improved stability. Its built-in i.f. amplifier stage, the gain of which is adjustable, permits use of the converter with receivers of widely-different performance characteristics.

Two r.f. stages are used, employing the highly-effective cascode circuit. The first tube is a 6AK5 connected as a triode, inductively neutralized. This feeds a 6J6 grounded-grid stage, in which one triode section is the amplifier, and all other tube elements are grounded. The mixer and oscillator are 6AB4 triodes. These functions could be combined in a single 12AT7 if desired, but separate triodes were used to permit more flexible adjustment of the oscillator injection. The mixer is followed by a 6AG5 i.f. amplifier, gain controlled by means of a potentiometer in its cathode circuit. The intermediate frequency is 7.4 Mc., selected because of its availability in most communications receivers, but 10.7 Mc., or any other desirable frequency, may be used.

The Oscillator

A high degree of receiver selectivity can be utilized effectively at 144 Mc. only if a stable and smooth-tuning oscillator is used in the converter. Mechanical vibration is reduced in this model through the use of a tank inductance made of ⅛-inch copper tubing, soldered directly to the stators of the tuning condenser. The latter is a type designed specifically for v.h.f. service. It has ball bearings at both ends of its rotor and ceramic end plates of heavy stock. Brackets for mounting the oscillator tube socket are an integral part of the condenser assembly. A smooth-operating dial assembly is made by substituting a large knob (National HRK or HRT) for the small one normally supplied with the National type K dial.

The oscillator circuit is one which provides constant output over the necessary tuning range, and the stage is run at low input, with light loading. The quality of the c.w. note thus obtained is adequate for reception of 2-meter c.w. signals, and the absence of hum modulation makes for good weak-signal reception of modulated signals. Oscillator injection is controlled by means of the link loops, L_{10} and L_{11}.

The R.F. and I.F. Stages

Though the converter has more tubes than the simplest units, it is not difficult to build or adjust. All circuits except the oscillator and the r.f. input circuit are slug-tuned, and only the oscillator is varied in tuning across the band. All stages may be peaked readily without a signal generator. The r.f. input circuit, L_2, is condenser-tuned, and it is important that a high-Q coil be used for best performance. The loading effect of the antenna is such that C_1 may be set for maximum signal at 146 Mc., and little difference in response will be noted at either end of the band.

The mixer and i.f. amplifier plate coils, L_6 and L_{7-8}, must be shielded, and coaxial line should be used for coupling the converter to the receiver, otherwise there may be annoying pick-up of signals at the intermediate frequency.

Construction

The position of components is not critical, and other arrangements may be desirable if the parts used are not duplicates of the original. In this instance an "L"-shaped layout is used, with the antenna terminals and r.f. stage at the right rear corner of the chassis and the second r.f., mixer, and i.f. amplifier stages running along the back and left sides in that order. The oscillator assembly is at the right

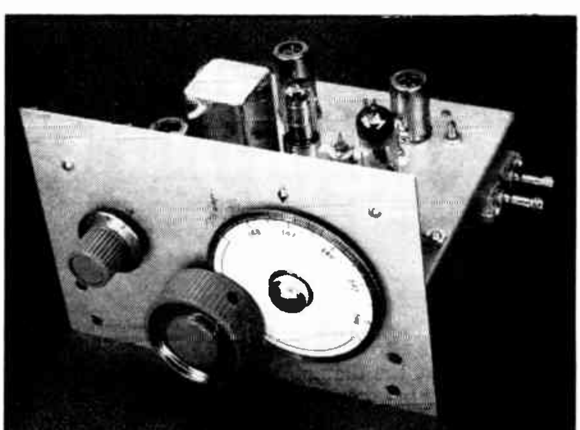


Fig. 12-6 — The cascode converter for 144 Mc. The dial calibration was made by drawing on heavy white paper, which is then fastened to the dial surface with rubber cement.

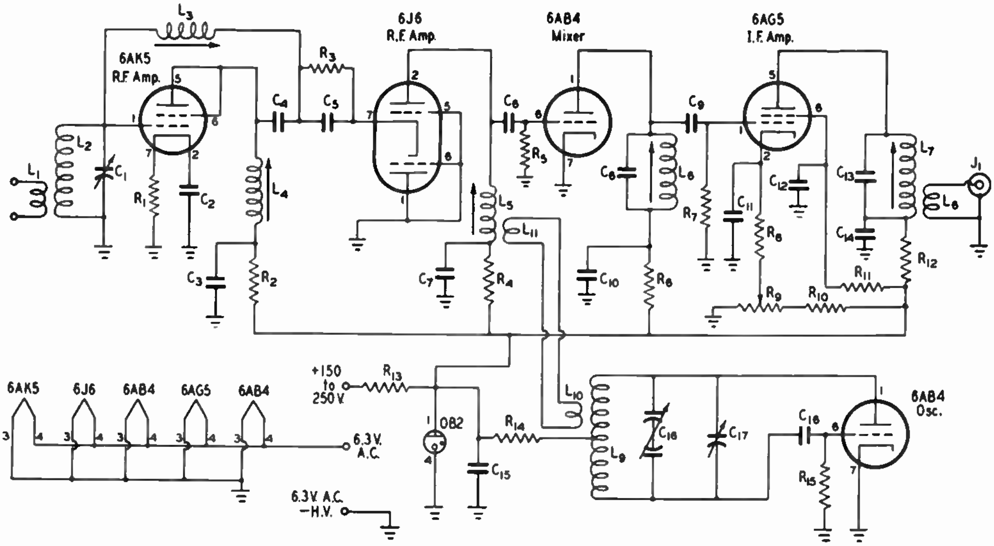


Fig. 12-7 — Schematic diagram of the 2-meter cascode converter.

C_1 — 8-μμfd. variable (Johnson 160-104).
C_2, C_3, C_7 — 470-μμfd. button-type by-pass.
C_4, C_6, C_8, C_{13}, C_{18} — 47-μμfd. ceramic.
C_5 — 470-μμfd. mica.
C_9 — 100-μμfd. ceramic.
C_{10}, C_{11}, C_{12}, C_{14} — 0.001-μfd. mica. (C_{10} and C_{14} are inside the i.f. shields.)
C_{15} — 75-μμfd. stand-off type by-pass.
C_{16} — 6.75-μμfd. stator-to-stator variable (National VHF-1-D).
C_{17} — 3–30-μμfd. air padder (Silver 619).
R_1, R_3, R_{14} — 100 ohms. (All resistors ½-watt unless otherwise specified.)
R_2, R_4, R_6, R_{12} — 1000 ohms.
R_5 — 0.68 megohm
R_7 — 1 megohm.
R_8 — 220 ohms.
R_9 — 2000-ohm wire-wound potentiometer.
R_{10} — 22,000 ohms, 1 watt.
R_{11} — 33,000 ohms.
R_{13} — 2500 ohms, 10 watts.
R_{15} — 15,000 ohms.
L_1 — 2 turns No. 18 enamel, ¾-inch diameter, between turns of L_2.
L_2 — 2 turns No. 14 tinned, ¾-inch diameter, ⅛ inch between turns.
L_3 — 10 turns No. 24 enamel on ¼-inch diameter slug-tuned form (CTC).
L_4, L_5 — 3 turns No. 24 enamel on ¼-inch diameter slug-tuned form (CTC). Winding ¼ inch long.
L_6, L_7 — No. 24 d.s.c.-wire close-wound to fill winding space on National XR-50 form.
L_8 — 5 turns No. 24 d.s.c. over cold end of L_7.
L_9 — Hairpin-shaped loop, ⅛-inch copper tubing, ¾ inch wide. Total length before soldering: 1½ inches. Extends 1⅛ inches beyond tuning-condenser stators. (See Fig. 12-9.)
L_{10}, L_{11} — Hairpin loops for coupling oscillator to mixer. See text and photographs.
J_1 — Coaxial connector.

front corner. It should be placed so that the flexible coupling does not touch the front panel. The chassis is aluminum, 7 by 7 by 2 inches, and the sheet aluminum panel measures 5¾ by 8 inches. Note that aluminum braces are used to prevent panel vibration. These were found necessary for best oscillator stability.

The method of coupling the output of the oscillator to the mixer may be seen in the bottom and rear views, Figs. 12-8 and 12-9. A coupling loop is mounted on the two outside lugs of a 3-lug tie-point strip directly below the oscillator inductance. This loop is connected through 75-ohm Twin-Lead to a loop around the r.f. plate coil, L_5. The center lug on the strip is used for mounting the oscillator decoupling resistor, R_{14}, which also serves as a third support point for the oscillator tank inductance. The size of the coupling loops, L_{10} and L_{11}, will depend on the amount of oscillator injection needed, but the degree of coupling will be small. L_{10} is a semicircular loop of No. 18 wire, ¾ inch across, about one-half inch below L_9. L_{11} is a circular loop concentric with L_5. It is visible in the left-hand corner of the bottom-view photograph, Fig. 12-8.

In mounting the oscillator tube socket the plate lug, pin No. 1, is soldered directly to the tuning-condenser stator. Pin No. 6 is connected to the other stator through the short length of the grid condenser, C_{18}. All other socket pins except the heater, No. 4, are connected together and grounded.

Adjustment

The first step in placing the converter in service is to set the oscillator for the proper frequency range, 136.6 to 140.6 Mc. for a 7.4-Mc. i.f. This may be done with a calibrated absorption-type frequency meter, or by listening to the oscillator on a calibrated receiver. Next the converter should be connected to the receiver with which it is to be used, and the i.f. adjustments (cores in L_6 and L_7) peaked for maximum noise. Next the slugs in L_4 and L_5 should be peaked for maximum noise, either tube noise or that from some external source, such as an electric razor or

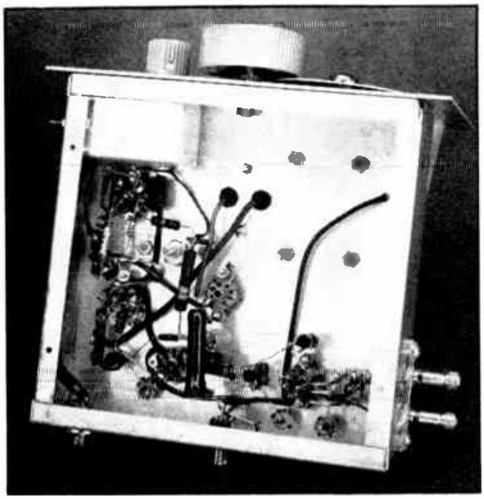


Fig. 12-8 — Bottom view of the 2-meter converter.

a noise generator. This should be done with the converter set for approximately 146 Mc. The r.f. input circuit may be peaked on noise or a signal by adjusting C_1, squeezing or spreading the turns of L_2 until the optimum setting occurs near minimum capacity. This adjustment should be made with the antenna connected.

Tuning of the slugs will be rather broad, so precise adjustment is not necessary. The slug in the neutralizing coil, L_3, may be set at approximately the midpoint of its travel, unless a noise generator is available, in which case it should be set for minimum noise figure. A noise generator will be helpful in determining the best position for L_1 with respect to L_2 also, but if none is available the coupling should be set somewhat *tighter* than that giving the maximum signal response.

The best position for the converter gain control will depend upon the sensitivity of the receiver with which the converter is to be used. With better-grade receivers it will be possible to operate the gain control well below the maximum setting. The optimum will be the minimum at which the over-all gain is adequate. The gain control also serves as a convenient means of setting up the S-meter reading, if the receiver is so equipped.

Coupling between the oscillator and mixer is not critical. The tighter the coupling the more the mixer output, within certain limits, but when an i.f. amplifier is used the highest possible mixer output is not required. The best setting of the coupling loop, L_{10}, is the *minimum* coupling required to give satisfactory response. Somewhat tighter coupling than the minimum required will have very little effect on the over-all performance, except to increase the pulling of the oscillator frequency as the second r.f. plate circuit is tuned. Very tight coupling will have an adverse effect on the signal-to-noise ratio and uniformity of response across the band.

A Simpler Version

If the builder desires the converter may be built in easy stages. In its simplest form it would consist only of the two 6AB4 stages, the mixer and oscillator. In this case the coil and condenser circuit, C_1L_2, would be substituted for the slug-tuned mixer coil, L_5, and the i.f. output would be taken off from the mixer plate coil, L_6. The i.f. amplifier stage should be added next, as it is quite essential to satisfactory operation. The addition of the r.f. stages provides a further improvement, particularly in signal-to-noise ratio in reception of weak signals.

The complete converter, as it is shown here, is the minimum that will provide performance sufficiently good to satisfy the discriminating v.h.f. worker, but the man who wishes to build something simpler as a start will be able to obtain reception of all but the weakest signals with the two- or three-tube version.

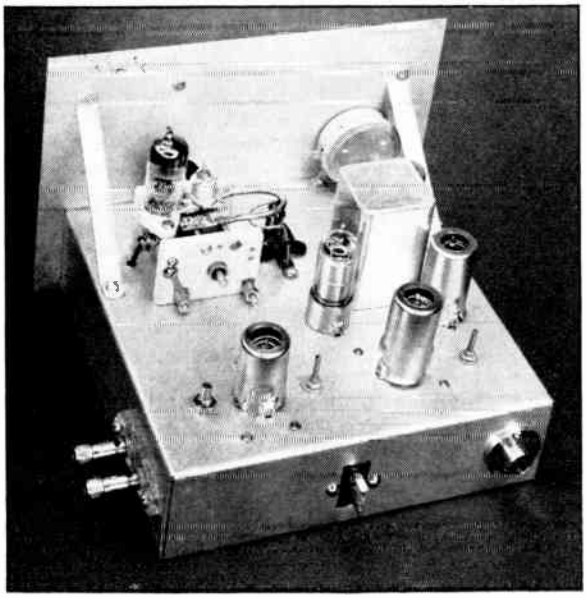


Fig. 12-9 — Rear view of the 2-meter converter. At the left side, near the panel, is the oscillator assembly. The r.f. stages, mixer, and i.f. amplifier are arranged in "L" formation across the back and right sides of the chassis, with the voltage-regulator tube in the middle.

A Crystal-Controlled Converter for 144 Mc.

Stability in the oscillator is of utmost importance in achieving satisfactory performance in a 2-meter converter, particularly when a highly-selective communications receiver is to be used as the i.f. amplifier. Even a small amount of drift, or the slightest mechanical instability, will make constant readjustment of the converter tuning necessary. In addition it is difficult, if not impossible, to secure freedom from hum modulation of incoming signals, when ordinary tunable oscillators are used. The converter shown in Figs. 12-10, 12-11 and 12-12 eliminates these difficulties by the use of a crystal-controlled oscillator at 13 Mc., followed by two multiplier stages to bring the frequency up to that required for 144-Mc. reception. Tuning of the band is accomplished by varying the communications receiver (now the i.f. amplifier) from 14 to 18 Mc.

At first thought it would appear that the design of such a converter would be complicated and its construction difficult, but the photographs and schematic diagram show that this is not necessarily true. Since no tunable circuits are required at the signal frequency the mechanical construction is simplified, and the alignment problems usually associated with tracking of gang-tuned circuits are reduced. Only four tubes are required, two of them r.f. amplifiers.

Circuit Details

Two 6J6s perform the functions of oscillator, multiplier and mixer. Two 6AK5s are used as bandpass r.f. stages, self-resonant overcoupled plate and grid circuits being employed to achieve the bandpass characteristics. Referring to the schematic diagram, Fig. 12-11, it will be seen that the crystal oscillator is a simple triode circuit, using one section of a 6J6 and a 13-Mc. crystal. The second section of the tube is a doubler, which drives the first section of the second 6J6 as a quintupler to 130 Mc. Energy from this stage mixes with the signal in the second section of the tube, the output of which is at the intermediate frequency.

Two problems are presented by this approach. First, the r.f. and mixer circuits must be broadened out sufficiently so that the response of the converter will be substantially flat over the entire band; and second, signals at the intermediate frequency may cause considerable interference unless the unit is completely shielded. The needed bandpass characteristics in the r.f. stages are supplied by overcoupling the stages, adjustment of which is explained in a later paragraph. The i.f. output transformer in the mixer plate circuit is provided with an adjustable tuned circuit, which requires some readjustment if maximum sensitivity is to be maintained across the entire band. It is not critical in its setting, however, so it does not complicate the tuning process appreciably.

The only other adjustment used after the initial tune-up procedure is completed is an r.f. gain control, the setting of which may be employed to reduce possible cross-modulation from extremely-strong local signals. Normally it may be set at the optimum position and left there without further change.

Mechanical Construction

Structural details should be reasonably clear from the photographs. The chassis is a "U"-shaped affair folded from sheet aluminum. Another folded sheet is used as a shield, completely enclosing the components, all of which are mounted on the main chassis, which is 2 by 5 by 6 inches in size. Looking at the top view, Fig. 12-10, it may be seen that the gain control and i.f. tuning adjustment are mounted on the front wall of the chassis. Across the top are the crystal and the two 6J6s, and at the rear are the antenna input terminals, the two 6AK5s, and the coaxial fitting for the i.f. output cable. The power plug is between the two r.f. tubes, at the rear edge of the chassis. Looking at the bottom of the chassis, Fig. 12-12, the oscillator, multiplier and mixer components are in the upper portion of the photograph, with the r.f. stages across the bottom.

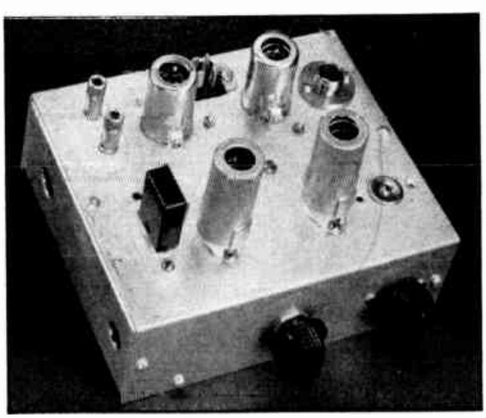


Fig. 12-10 — A crystal-controlled converter for 144 Mc. Two 6J6s serve as oscillator, multiplier and mixer. The two 6AK5s are bandpass r.f. stages.

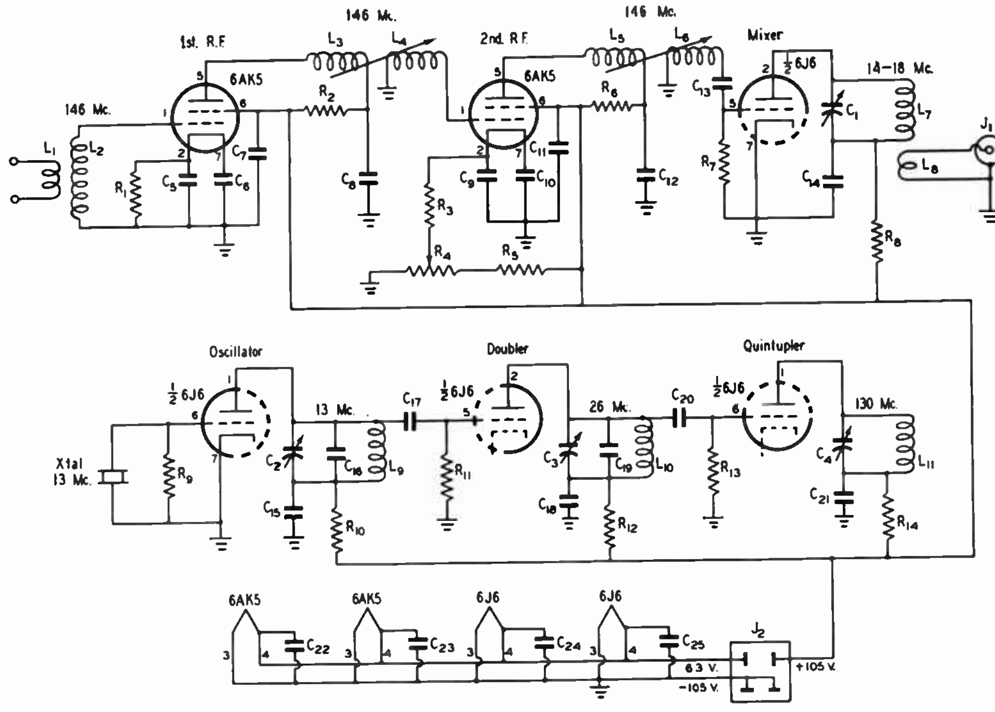


Fig. 12-11 — Wiring diagram of the 2-meter converter with crystal-controlled oscillator.

C_1 — 15-μμfd. variable (Millen 20015).
C_2, C_3, C_4 — 3–30-μμfd. mica trimmer.
C_5, C_6, C_7, C_9, C_{10}, C_{11}, C_{21}, C_{22}, C_{23}, C_{24}, C_{25} — 250-μμfd. ceramic.
C_8, C_{12} — 75-μμfd. ceramic.
C_{13} — 50-μμfd. ceramic.
C_{14}, C_{15} — 0.01-μfd. ceramic.
C_{16}, C_{17}, C_{20} — 100-μμfd. ceramic.
C_{18} — 0.0015-μfd. ceramic.
C_{19} — 27-μμfd. ceramic.
R_1, R_3 — 220 ohms, ½ watt.
R_2, R_6, R_8, R_{10}, R_{12}, R_{14} — 1000 ohms, ½ watt.
R_4 — 2000-ohm wire-wound potentiometer.
R_5 — 10,000 ohms, 1 watt.
R_7 — 8.2 megohms, ½ watt.
R_9 — 22,000 ohms, ½ watt.
R_{11}, R_{13} — 0.1 megohm, ½ watt.
L_1 — 2 turns No. 18 enameled wire, ⅜-inch inside diameter, tightly coupled to ground end of L_2.
L_2, L_4, L_6, L_{11} — 5 turns No. 14 tinned wire, ⅜-inch diameter, turns spaced wire diameter (see text).
L_3 — 6 turns No. 14 tinned wire, ⅜-inch diameter, turns spaced wire diameter (see text).
L_5 — 7 turns No. 14 tinned wire, ⅜-inch diameter, turns spaced wire diameter (see text).
L_7 — 27 turns No 28 enameled wire, close-wound to a length of ⅜ inch on a ½-inch diameter form.
L_8 — 3 turns "push-back" wire, close-wound over cold end of L_7.
L_9 — 8 turns No. 18 tinned wire, ¾-inch diameter, ½ inch long.
L_{10} — 6 turns No. 18 tinned wire, ¾-inch diameter, ⅜ inch long.
Note: Coils L_9 and L_{10} are cut from a length of B & W "Miniductor" type 3011.
J_1 — Coaxial-cable connector (Jones S-201).
J_2 — 4-prong male plug (Jones P-304-AB).

The shield arrangement which isolates the r.f. stages may also be seen in the bottom view. It is made of sheet copper, to permit easy soldering to the cylindrical shields on the tube sockets. At the left is the grid coil, L_2, with the antenna winding, L_1, inserted between its first two turns. Inside the shield are the first r.f. plate (L_3) and second r.f. grid (L_4) coils mounted close together in the same plane. Just outside the upper right corner of the shield are the second r.f. plate coil, L_5, and the mixer grid coil, L_6.

At the upper left of the bottom view are the oscillator and doubler plate coils, L_9 and L_{10}, with their associated trimmers. The quintupler plate coil and its trimmer are between the gain control and the i.f. output tuning condenser. The i.f. output transformer, L_7L_8, is at the upper right. Because of their small size and greater effectiveness, ceramic capacitors were used throughout the unit, the particular values used now being available from Centralab and possibly others.

Adjustment and Operation

The first step in putting the converter into service is to set up the oscillator and multiplier stages for proper operation. This procedure is similar to that employed in multistage transmitters, except that at 105 volts the currents in the various stages are very low. It is recommended that the supply voltage be maintained at that figure with a VR-105, in which case the total drain for the two 6J6s is about 12 ma. The two 6AK5s draw approximately the same. A low-range milliammeter may be inserted in series with the plate decoupling resistors, R_{10}, R_{12} and R_{14}, to check on the operation of the oscillator and multiplier stages, as a good plate-current dip is noted as the 6J6 stages are tuned to resonance. A calibrated absorption-type wavemeter should be used to be certain that the doubler and quintupler stages are operating on the correct frequencies.

The r.f. circuits may be aligned by means of a grid-dip meter. Details of a suitable instrument, which is a useful tool for many other purposes, will be found in Chapter Sixteen. The inductance of the grid and plate coils may be adjusted to approximately the proper value by spreading or squeezing the turns to the point which produces maximum dip with the grid-dip meter set for 146 Mc.

The mixer plate circuit may be checked by feeding a signal into the mixer grid at the i.f. frequency, 14 to 18 Mc., making sure that the tuned circuit, C_1L_7, is capable of resonating across this range. In the absence of a signal generator, 20-meter amateur signals may be used for this check by connecting an antenna to the mixer grid. They should peak with the mixer plate condenser set near the middle of its tuning range.

A calibrated signal generator operating in the 144-Mc. range is helpful in checking the converter performance, but a low-powered oscillator, or even a superregenerative receiver, may be used. A fair idea of the performance can be obtained by merely connecting an antenna to the converter and aligning the r.f. circuits by noise pick-up. If the antenna in question is a simple dipole cut for 146 Mc., the noise level should remain nearly constant over the entire band, if the r.f. coils and the coupling between them has been adjusted properly. If necessary, the self-resonant coils may be "stagger-tuned" to achieve uniform response across the band.

Other Suggested Circuits

One of the problems in connection with the use of crystal-controlled oscillators in v.h.f. converters is the choice of a suitable crystal frequency. If a relatively low frequency is used in the crystal oscillator the crystal must be chosen carefully to avoid trouble from its harmonics falling in the band to be covered. If higher crystal frequencies are used the cost of the crystal becomes considerable, and some harmonic-type crystals have poor stability.

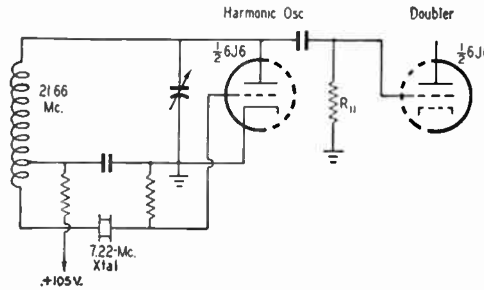


Fig. 12-13 — Schematic diagram of a regenerative harmonic oscillator circuit, which may be substituted for the 13-Mc. oscillator shown in Fig. 12-11. The second section then doubles to 43.32 Mc., and the second 6J6 operates as a tripler to 130 Mc., instead of a quintupler.

A crystal oscillator circuit which helps to by-pass these troubles is described in the transmitter chapter in connection with the v.h.f. exciter-transmitter pictured in Fig. 13-11. With this circuit, Fig. 13-12, ordinary 7–8-Mc. crystals are made to oscillate on their third harmonic, thus reducing the number of stages required, and permitting the use of inexpensive crystals.

This circuit may be substituted for that in Fig. 12-11, in case a less expensive or more readily-obtainable crystal is to be used. An example would be the use of a 7.22-Mc. crystal, oscillating at 21.66 Mc. in the first 6J6 triode section of Fig. 12-11. The second section would double to 43.32 Mc. The next triode section would operate as a tripler to 130 Mc. Except for the grid circuit of the first 6J6 the schematic would be similar to that shown in Fig. 12-11. The regenerative harmonic oscillator circuit is reproduced in Fig. 12-13. It is suggested that prospective users study the material on page 421, Chapter Thirteen, for further information before attempting to utilize this circuit for receiver purposes.

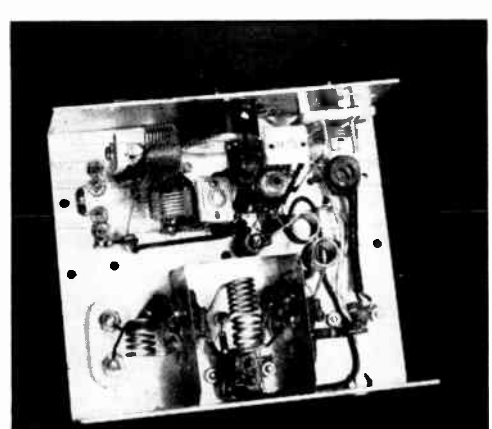


Fig. 12-12 — Bottom view of the crystal-controlled converter. Note the overcoupled circuits in the two r.f. stages.

A Cascode Converter for 220 Mc.

The 220-Mc. converter shown in Figs. 12-14, 12-15 and 12-16 is an adaptation of the 2-meter design shown earlier in this chapter. The cascode r.f. amplifier is similar, but different tubes are used in the mixer-oscillator and i.f. amplifier stages, and a completely different mechanical layout is employed, as is quickly evident from a comparison of the photographs.

The first stage is a triode-connected 6AK5. No neutralization was found necessary, oscillation being prevented by the heavy loading imposed by the following 6J6 grounded-grid stage. The r.f. plate circuits are self-resonant and sufficiently broadband to more than cover the 220-Mc. band without adjustment. The r.f. input circuit is condenser-tuned, but the antenna loading makes repeaking of this circuit unnecessary.

The functions of mixer and oscillator are combined in a 12AT7 dual triode. The oscillator tank circuit is in the form of a "U" cut from sheet copper and soldered directly to the tuning-condenser stators. A fairly high value of parallel capacitance is added in C_4, for stability. Some mixer injection is obtained through the elements and common connections of the dual triode, but additional coupling was found to be necessary. It is added by a short piece of Twin-Lead, and shown on the diagram as C_{10}.

The 6BA6 i.f. amplifier has a potentiometer in its cathode circuit for gain control. The mixer plate coil, L_5, and the output transformer, L_6, make use of ready-made commercial slug-tuned coils of small dimensions, requiring only slight modification of L_6, as noted in the parts list. Voltages throughout the converter are stabilized by an 0B2 regulator tube.

Mechanical Details

As may be seen in the photographs, the r.f. and mixer-oscillator tubes are mounted in an inverted position, with the sockets above the chassis. This keeps r.f. leads to a minimum (of great importance in 220-Mc. construction) and brings portions of the circuit requiring adjustment up where they are readily accessible. The i.f. and voltage-regulator tubes are mounted in the conventional manner. The i.f. tuning slugs are below the chassis, providing partial shielding. If pick-up of signals on the i.f. frequency is troublesome it can be corrected by the addition of a bottom plate to the chassis.

An aluminum chassis, 2 by 5 by 7 inches, is used, with a 5 by 7-inch panel. A small bracket is mounted at the back of the chassis, carrying the r.f. tuning condenser and the crystal socket used for antenna terminals. The antenna condenser is insulated from ground at the mounting point, and a heavy copper strip is run over to the common ground point at the r.f. tube socket. Ceramic fixed condensers (the disk type in the higher values and the cylindrical type in the smaller ones) allow compact design, and provide improved by-passing qualities.

Alignment

Putting the converter into service involves only standard procedure, such as that outlined in the description of the 2-meter converter of similar design, except that the plate coils, L_3 and L_4, are adjusted by spacing their turns. The intermediate frequency can be anything within reason, depending on the receiver with which the converter is used. In the original model it is 15 Mc., but it could be altered considerably without component change, other

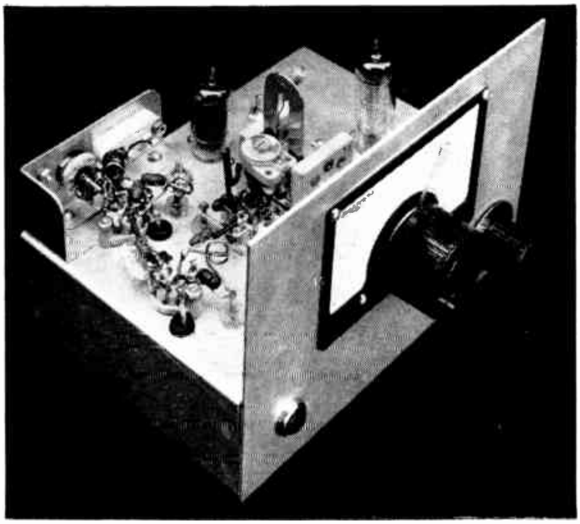


Fig. 12-14 — Oblique view of the 220-Mc. cascode converter, showing inverted mounting of the r.f. and mixer-oscillator tubes. The r.f. input circuit is mounted on a bracket at the rear of the chassis, with the 6AK5 socket directly below it. Nearer the panel is the 6J6, with the 12AT7 mixer-oscillator socket and oscillator components near the middle. The 6BA6 i.f. amplifier and 0B2 regulator tubes are mounted in the conventional manner, at the right.

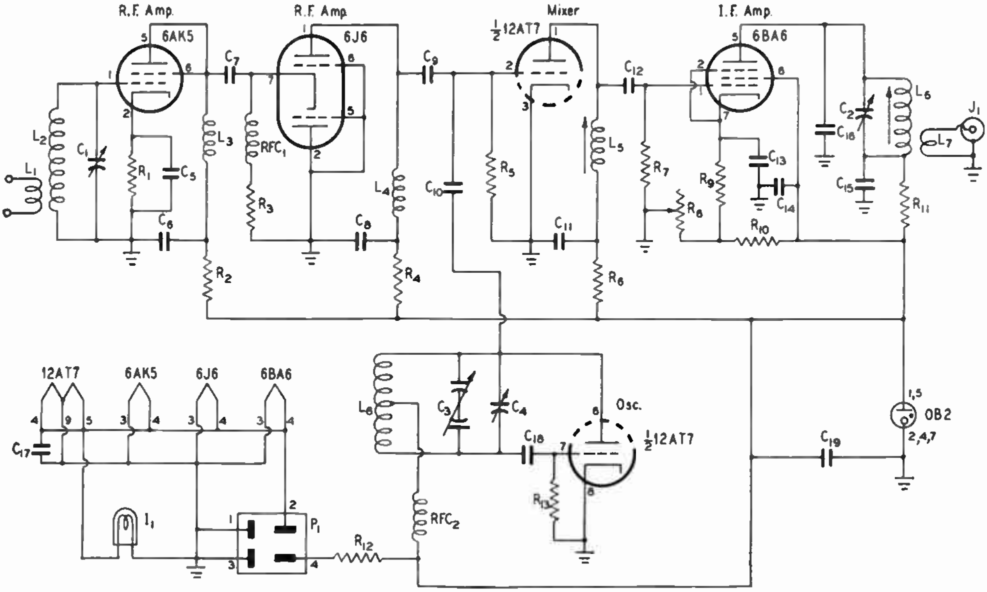


Fig. 12-15 — Schematic diagram of the 220-Mc. converter.

C_1 — 5-$\mu\mu$fd. miniature variable (Johnson 160-102).
C_2 — 3–30-$\mu\mu$fd. mica trimmer.
C_3 — 10-$\mu\mu$fd.-per-section split stator (Bud LC-1660 with 2 rotor plates removed from each section).
C_4 — 5–25-$\mu\mu$fd. ceramic trimmer.
C_5, C_6, C_8, — 100-$\mu\mu$fd. ceramic tubular.
C_7, C_9, C_{12}, C_{16}, C_{18} — 25-$\mu\mu$fd. ceramic tubular.
C_{10} — 75-ohm Twin-Lead, cut to approximately ½ inch.
C_{11}, C_{13}, C_{14}, C_{15}, C_{17} — 0.001-μfd. disk-type ceramic.
C_{19} — 0.01-μfd. disk-type ceramic.
R_1, R_9 — 68 ohms, ½ watt.
R_2, R_4, R_6, R_{11} — 1000 ohms, ½ watt.
R_3 — 100 ohms, ½ watt.
R_5, R_7 — 1 megohm, ½ watt.
R_8 — 25,000-ohm potentiometer.
R_{10} — 56,000 ohms, 1 watt.
R_{12} — 5000 ohms, 10 watts.
R_{13} — 12,000 ohms, ½ watt.
L_1 — 3 turns No. 18 enamel, ½-inch diameter, inserted between turns of L_2.
L_2 — 3 turns No. 16 tinned, ½-inch diameter, ⅜ inch long.
L_3, L_4 — 2 turns No. 16 tinned, ¼-inch diameter. Space turns for maximum response.
L_5 — I.f. plate coil, slug-tuned (CTC LS3-10 Mc.).
L_6 — Similar to L_5, but with 10 turns removed.
L_7 — 10 turns No. 30 d.s.c. wound over L_6.
L_8 — Copper loop, 1/16 inch thick, ⅞ inch wide, 1⅝ inches long. Slot is ½ inch wide.
I_1 — Green-jewel pilot-lamp assembly.
J_1 — Coaxial output fitting.
P_1 — 4-prong male plug.
RFC_1, RFC_2 — 0.84-μh. r.f. choke (Ohmite Z-235).

than different settings of the slugs in L_5 and L_6, and resetting of padder C_4, across the oscillator circuit. It will be noted that part of the capacitance across L_6 is in the form of a fixed condenser, C_{16}. This is to forestall possible v.h.f. parasitic oscillation in the i.f. stage, by by-passing the plate direct to the cathode. Tunable capacitance, C_2, is connected across the i.f. coil. Its adjusting screw is reached through a rubber-grommetted hole in the side of the chassis.

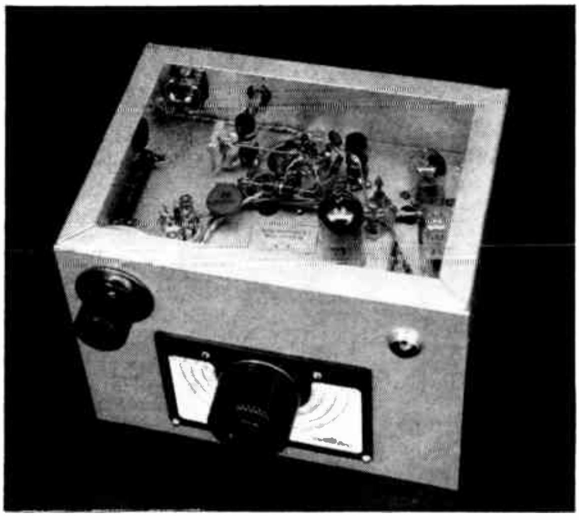


Fig. 12-16 — Under chassis view of the 220-Mc. converter. The three inverted tubes may be seen at the right. The slug-tuned i.f. coils are at the back of the chassis.

The Superregenerative Receiver

The simplest type of v.h.f. receiver is the superregenerator, for many years the most popular receiver for v.h.f. work. It affords fair sensitivity with few tubes and elementary circuits, and though it has largely been replaced by the more effective superheterodyne for home-station use, it still has many v.h.f. applications. Its disadvantages are lack of selectivity, poor signal-to-noise ratio on weak signals, and its tendency to radiate a strong signal which causes severe interference.

Its selectivity may be improved somewhat and its interference capabilities reduced by the addition of an r.f. stage, a refinement which should be considered a necessity if the receiver is to be used in a locality where there are other stations operating on the same band. If no r.f. stage is used, as in portable applications where economy of space and battery drain are primary considerations, the detector should be operated with the lowest plate voltage that will permit superregeneration, in order to reduce its interference range.

From a practical aspect, superregenerative receivers may be divided into two general types. In the first the quenching voltage is developed by the detector itself, called a "self-quenched" detector. In the second, a separate low-frequency oscillator is used to generate the quench voltage. Self-quenched detectors have found wide favor, particularly for portable work; but it is possible to achieve better performance with the separately-quenched type, particularly as the frequency approaches the upper limit of the tube's capabilities.

Superregeneration Principles

The limit to which ordinary regenerative amplification can be carried is the point at which oscillation commences, since at that point further amplification ceases. The superregenerative detector overcomes this limitation by introducing into the detector circuit an alternating voltage of a frequency somewhat above the audible range, the value being between 20 and 200 kc. depending on the signal frequency. Because the oscillations are constantly being interrupted by this quenching voltage the regeneration can be greatly increased, and the amplified signal will build up to tremendous proportions. A one-tube superregenerative receiver is capable of an inherent sensitivity approaching the thermal-agitation noise level of the tuned circuit, and may have an antenna input sensitivity of two microvolts or better.

Because of its inherent characteristics, the superregenerative circuit is suitable only for the reception of modulated signals, and operates best on the very-high frequencies. Typical superregenerative circuits for separately-quenched and self-quenched detectors are shown in Fig. 12-17, but the basic circuit may be any of the various arrangements used for straight regenerative detectors.

In the self-quenched detector the frequency of the quench oscillation depends upon the feed-back and upon the time constant of the grid leak and condenser, the oscillation being a "blocking" or "squegging" in which the grid accumulates a strong negative charge which does not leak off rapidly enough through the grid leak to prevent a relatively slow variation of the operating point.

The greater the difference between the quenching and signal frequencies the greater the amplification, because the signal then has a longer period in which to build up during the nonquenching half-cycle when the resistance of the circuit is negative. This ratio should not exceed a certain limit, however, for during the quenched or nonregenerative intervals the input selectivity is merely that of the Q of the tuned circuit alone.

Because of the greater amplification, the hiss noise when a super-

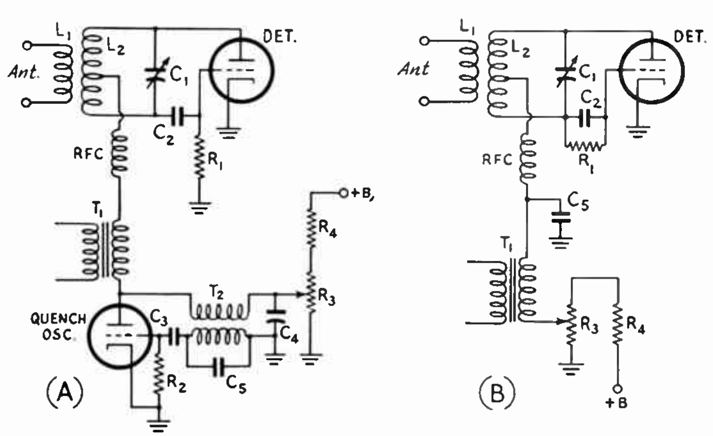


Fig. 12-17 — (A) Superregenerative detector circuit using a separate quench oscillator. (B) Self-quenched superregenerative detector circuit. L_2C_1 is tuned to the signal frequency. Typical values for other components are:

C_2 — 47 μμfd.
C_3 — 470 μμfd.
C_4 — 0.1 μfd.
C_5 — 0.001–0.047 μfd.
R_1 — 2–10 megohms.
R_2 — 47,000 ohms.
R_3 — 50,000-ohm potentiometer.
R_4 — 47,000 ohms.
RFC — R.f. choke, value depending upon frequency. Small low-capacitance chokes are required for v.h.f. operation.
T_1 — Audio transformer, plate-to-grid type.
T_2 — Quench-oscillator transformer.

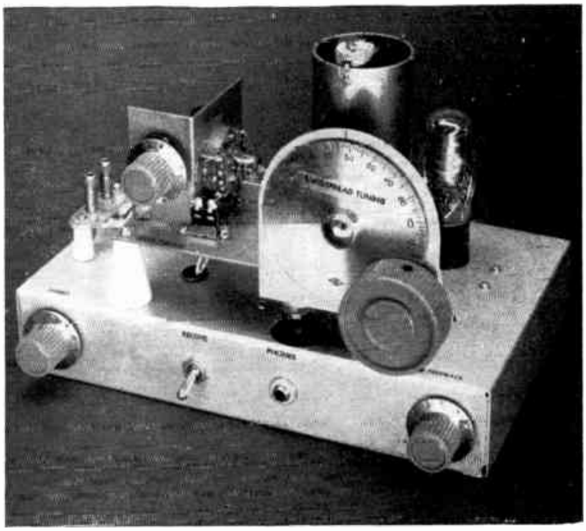


Fig. 12-18 — Front view of the coaxial-line receiver. The r.f. amplifier tuning control is at the left and the main control, for the concentric-line detector circuit, is at the right side of the unit. The audio gain control, send-receive switch, 'phone jack and regeneration control can be seen in that order, from left to right, across the front wall of the chassis.

regenerative detector goes into oscillation is much stronger than with the ordinary regenerative detector. The most sensitive condition is at the point where the hiss first becomes marked. When a signal is tuned in, the hiss will disappear to a degree that depends upon the signal strength.

Lack of hiss indicates insufficient feed-back at the signal frequency, or inadequate quench voltage. Antenna-loading effects will cause dead spots that are similar to those in regenerative detectors and can be overcome by the same methods. The self-quenching detector may require critical adjustment of the grid-leak and grid-condenser values for smooth operation, since these determine the frequency and amplitude of the quench voltage.

A COAXIAL-LINE SUPERREGENERATIVE RECEIVER FOR 220 MC.

The performance of a superregenerative receiver, both as to selectivity and smoothness of operation, can be improved by the use of a coaxial line tank in the grid circuit of the detector, in place of the customary coil and condenser. Addition of an r.f. amplifier stage will improve sensitivity, reduce radiation, and make antenna coupling less critical. A superregenerative receiver for 220 to 240 Mc. incorporating these features is shown in Figs. 12-18–12-21.

The r.f. tube is a 954 acorn with a conventional tuned circuit in its grid. The plate circuit is a self-resonant loop, which is coupled to the concentric line grid circuit of the 6AK5 detector. The detector output is fed through a quench filter to a

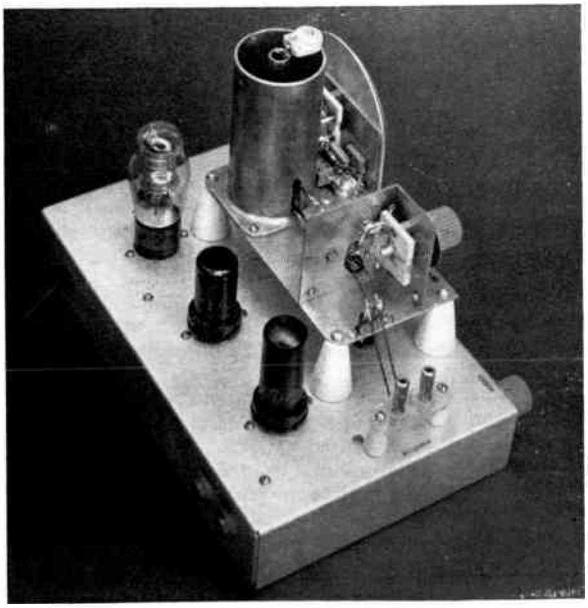


Fig. 12-19 — Rear view of the superregenerative receiver. The r.f. circuits are mounted on a copper shelf to the left of the antenna terminals. The detector tuning condenser is mounted on a small panel to the front of the coaxial line, and the band-set condenser is soldered across the open end of the line. The r.f. stage is mounted on an "L"-shaped bracket with the tube socket and plate-circuit components on the right side and the grid circuit on the right side. Audio tubes and voltage regulator are in line across the rear of the chassis.

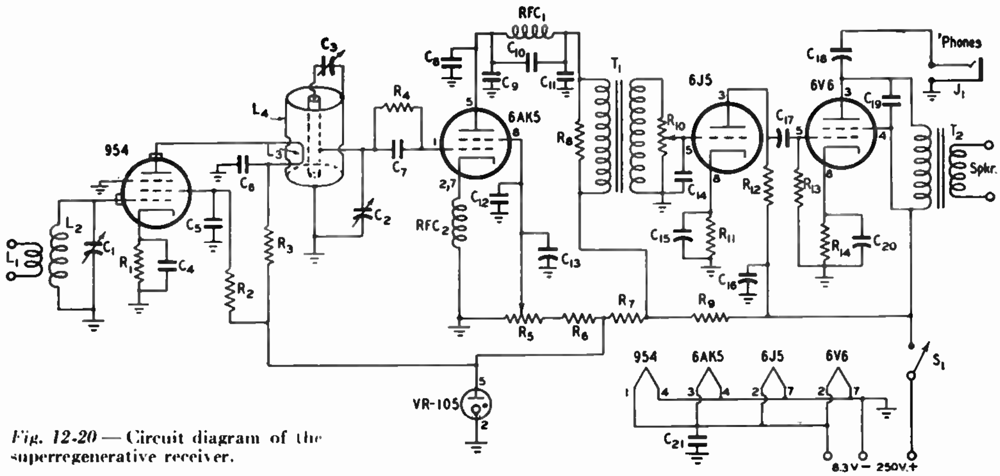


Fig. 12-20 — Circuit diagram of the superregenerative receiver.

C_1 — Midget variable condenser (Millen 20015 reduced to one stator and two rotor plates).
C_2 — Midget variable condenser (Millen 20015 reduced to one stator and one rotor plate).
C_3 — 5–20-μμfd. ceramic trimmer (Centralab 820-B).
C_4, C_5, C_6 — 100-μμfd. (National XLA-C).
C_7 — 22-μμfd. mica.
C_8, C_{12}, C_{21} — 470-μμfd. mica.
C_9 — 0.0022-μfd. mica.
C_{10}, C_{11} — 0.0068-μfd. mica.
C_{13} — 0.2-μfd. 400-volt paper.
C_{14} — 47-μμfd. mica.
C_{15} — 10-μfd. 25-volt electrolytic.
C_{16} — 8-μfd. 450-volt electrolytic.
C_{17}, C_{18} — 0.01-μfd. 400-volt paper.
C_{19} — 0.0047-μfd. mica.
C_{20} — 100-μfd. 25-volt electrolytic.
R_1, R_3 — 1000 ohms, ½ watt.
R_2 — 33,000 ohms, ½ watt.
R_4 — 0.1 megohm, ½ watt.
R_5 — 50,000-ohm potentiometer.
R_6 — 47,000 ohms, 1 watt.
R_7, R_9 — 1500 ohms, 10 watts.
R_8 — 22,000 ohms, ½ watt.
R_{10} — 0.25-megohm potentiometer.
R_{11} — 2200 ohms, ½ watt.
R_{12} — 0.1 megohm, ½ watt.
R_{13} — 0.47 megohm, ½ watt.
R_{14} — 270 ohms, 1 watt.
L_1 — 2 turns No. 18 e., ¼-inch inside diameter, close-wound.
L_2 — 2 turns No. 12 e., ¼-inch inside diameter, ⅛-inch space between turns.
L_3 — 5¼-inch length No. 12 e., bent to form a "U"-shaped loop having a ¾-inch space between conductors. Plate side of loop is 1¾ inches long and the opposite side is 2¾ inches long.
L_4 — Concentric line. Inside conductor is a 4-inch length of ½-inch o.d. copper tubing. Grid tap 1 inch from grounded end for both 220- and 235-Mc. operation or ¾ inch from grounded end for 220 Mc. only. Outside conductor is a 4-inch length of 2-inch i.d. copper tubing.
J_1 — Open-circuit jack.
RFC_1 — 80-mh. choke (Meissner 19-5596).
RFC_2 — 1-mh. r.f. choke (National R-33).
S_1 — S.p.s.t. toggle switch.
T_1 — Interstage audio transformer (Stancor A-53C).
T_2 — Universal output transformer (Cinaudagraph U-85).

6J5 triode audio followed by a 6V6 second audio. Either 'phones or 'speaker may be used.

Constructional Details

The receiver is built on a standard aluminum chassis measuring 2 by 7 by 11 inches and the small panel for the detector tuning dial is cut from a sheet of $\frac{1}{16}$-inch aluminum measuring 3⅞ by 3⅞ inches. The shelf for the r.f. section is made from a piece of $\frac{1}{16}$-inch copper stock measuring 5½ by 6¼ inches which is cut and bent as shown in the photographs of the receiver. The horizontal section of the subchassis measures 3½ by 6¼ inches and the small vertical panel is 2 inches high and 2½ inches wide. The detector bandspread condenser and the aluminum panel for the detector tuning dial are both mounted on this upright member of the copper chassis. C_2 is mounted with the two stator terminals facing toward the right end of the chassis (as seen from the rear view) and the lower stator terminal is one inch up from the horizontal surface and 1¼ inches in from the left side of the copper panel. The tube socket for the 6AK5 is 2 inches in from the left end of the chassis and is located as far toward the front edge as possible.

The "L"-shaped bracket for the r.f. amplifier is 2½ inches high, has a depth of 2⅜ inches, and is 1½ inches across the front. Spade lugs are bolted, and then soldered, to the bottom of the partition to provide a method of mounting that is both electrically and mechanically sound. The National XLA tube socket is centered on the side of the partition at a point located 1⅜ inches in from the rear and top edges. A $\frac{5}{16}$-inch hole, drilled in the bracket at this point, allows the grid prong of the 954 to extend through to the grid-circuit components. The cathode and heater prongs of the socket face toward the front of the receiver and the XLA-C by-pass condensers are mounted inside the socket. The plate by-pass condenser, C_6, is mounted underneath socket prong No. 5 as this prong is used as the support point for the cold end of the plate loop, L_3. Note that the No. 5 prong is a spare so far as the 954 is concerned. A National XLA-S internal shield, designed for use with the XLA socket, provides a common path for the condenser ground connections and, of course, this soldering should be done before the socket is bolted to the copper partition. The heater, cathode and suppressor connections are also made to the internal shield and, after mounting, the shield is in turn soldered to the copper plate.

The r.f. amplifier tuning condenser is mounted with the shaft in line with the shaft of C_2. Stator terminals face to the left so that the bottom terminal is within ¼ inch of the 951 grid prong. L_2 is supported by the condenser terminals and the antenna coil, L_1, is supported by L_2 and by the two-terminal lug strip located to the right of the amplifier. Grid clips for the 954 were improvised by removing the prongs from a miniature tube socket.

Holes, large enough to clear 6/32 machine screws, are drilled at each corner of the copper mounting plate so that the unit may be mounted on 1½-inch stand-off insulators. Larger holes, equipped with rubber grommets, are adjacent to the detector and amplifier tube sockets so that power wiring may be passed down through the main chassis.

Construction of the concentric line is not difficult if the various operations are carried out as suggested below. The inner and outer conductors are 4 inches long, and the end plate is 2½ inches square. A ½-inch hole for the inner conductor of the line should be drilled at the center of the end plate, and the plate should also have a hole for a 6/32 machine screw at each corner. However, before the center hole is drilled, it is advisable to use the center-punch mark as the pivot for scribing a circle to indicate the position of the outside conductor. This will simplify the task of lining up the two pipes for the soldering operation.

A 3/8-inch hole should now be drilled in the large pipe at a point located 1 inch up from the bottom edge, and a second hole of 5/16-inch diameter should be drilled on line with the larger hole and around the pipe by 90 degrees. After the material between these two holes and the bottom of the tubing is removed by cutting with a hack saw, the finished slots will provide openings for the input coupling coil, L_3, and the detector-grid connection. The inner conductor should also be drilled and tapped for a 6/32 machine screw at this time. One hole, ¾ inch up from the bottom of the line, is required if the receiver is to be used to cover only one band. A second hole, ¼ inch above the first, is necessary if the receiver is to be tuned to both the 220- and 235-Mc.* bands. In either case, the tapped hole will be used as the connecting point for the lead running to the tuning condenser.

Unless extremely thin-walled tubing is used for the concentric line, it will be difficult to complete the soldering operation with an ordinary iron. Placing the assembly on an electric hot plate will heat the copper in a very few minutes and will allow the work to be done neatly and easily. The end plate should be laid on a flat level surface while the inner conductor is lined up perpendicular to the horizontal surface of the plate. This operation may be carried out with the metal resting on the hot plate if the latter is to be used. The outer conductor should be placed in the position indicated by the scribed circle. Heat may now be applied and the soldering completed. The metal is ready to accept solder when a rapid change in the color of the copper is noticed. A long piece of solder may be inserted through the open end of the line, and as the end is moved around the surfaces to be joined the solder will melt and run into place easily.

The remaining constructional work is straightforward and study of the three photographs will show the location of the various components. Since there is no crowding of parts, it should not be difficult to duplicate the original layout.

*The 235-Mc. band is still assigned in Canada at this writing.

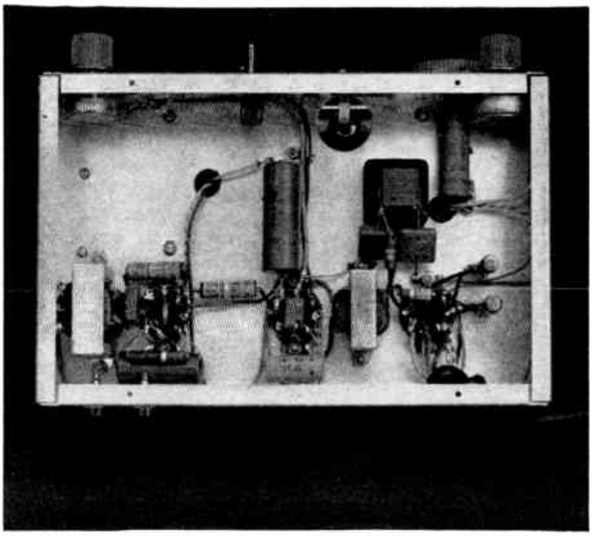


Fig. 12-21 — Bottom view of the coaxial-line receiver showing the output transformer, T_2, located at the lower left-hand corner of the chassis, and the audio transformer, T_1, mounted between the sockets for the audio tubes. The quench-filter choke, housed in a metal shield, is above and to the right of T_1. Resistors R_7 and R_6 are mounted on end to the right of the regulator-tube socket.

A Mobile Converter for 28 and 50 Mc.

The converter shown in Figs. 12-22–12-25 was designed for mobile reception on 6, 10, and 11 meters, but it may also be used in fixed-station work with good results. The intermediate frequency is 1500 kc., to permit its use with mobile broadcast receivers.

Circuit Details

The converter circuit diagram is shown in Fig. 12-23. A 6AK5 broadband r.f. amplifier is followed by a 6J6 mixer-oscillator. The oscillator circuit is the ultraudion type, operating 1500 kc. below the signal frequency. The need for gang-tuned circuits is eliminated by the broadband r.f. amplifier; thus only the oscillator tuning condenser, C_1, requires adjustment during normal tuning operation. Band changing is accomplished with a 5-section selector switch, shown on the diagram as $S_{1A, B, C, D, E}$.

Fig. 12-22 — A bandswitching converter for 6, 10 and 11 meters. The pilot light at the lower right has an adjustable beam, for convenience in mobile work.

Seven commercially-available coils are used, six of them being identical except for the setting of the slugs. The wide inductance range of the slug-tuned units makes it possible to use similar coils for the r.f., mixer and oscillator coils for both ranges. Padder capacitance is added across the 10-meter r.f. and mixer coils, L_4 and L_6, and across both oscillator coils, L_7 and L_8. Varying the slug position takes care of the necessary differences in coil inductance for all these positions.

A single whip antenna may be used for both broadcast and amateur reception. A jumper connection between sections *A* and *E* of S_1 completes the circuit between the antenna and the broadcast receiver, with the switch in the position marked *B.C.* on Fig. 12-23. A filament switch, S_2, is provided to remove the load of the converter tubes from the car battery when the receiver is being used for broadcast reception.

Broadbanding of the r.f. and mixer circuits is accomplished through the use of low-*Q* coils and tight coupling in the antenna circuit. The plate coil of the mixer is self-resonant at the i.f. frequency, giving a degree of broadness sufficient to permit tuning the receiver over a limited range near the high end of the broadcast band, providing a vernier effect.

Construction

All of the metal components are formed from 1/16-inch aluminum stock. The interior view, Fig. 12-24, shows the "L"-shaped section which serves as the front panel and the bottom plate of the unit. The panel and the bottom areas are each 5 inches square. Lips, 1/2 inch wide, are folded over along the top and side edges of the panel and also along the sides of the bottom section. The rolled-over edges are drilled and tapped to accommodate 6-32 machine screws.

A three-sided portion and a square top plate complete the converter cabinet. The sides are 5 inches square and the rear wall is 5 1/8 inches wide. All three sides are 5 inches high with 1/2-inch flanges folded over on the top edges and drilled and tapped for 6-32 screws. The sides and bottom edges of the case are drilled to clear machine screws; the holes should line up with the tapped holes of the panel-bottom assembly. A rectangular hole, 1 7/8 inches high and 2 inches wide, is cut at the bottom left-hand corner (as seen from the rear of the converter) of the rear wall, to provide clearance for the cable connectors. The top plate for the converter measures 5 by 5 inches. Holes, drilled along the edges, allow the cover to be fastened to the flanges at the top of the cabinet.

The physical shape of the converter chassis can best be visualized by study of the interior views. The chassis is 5 by 4 7/8 by 1 3/4 inches in size, with flanges 1/2 inch wide folded over along the front and the bottom edges to provide a means of mounting. A 2 1/4 × 3 3/4-inch cut-out at the center of the chassis allows clearance for the bandswitch. A large round hole located in the rear wall of the chassis simplifies the job of finding the oscillator padder condenser when this control requires adjustment.

A vertical partition used as the mounting surface for the oscillator tuning condenser, C_1, also serves as the shield between the plate and the grid circuits of the r.f. amplifier. It is 3 1/2 inches wide and 4 3/4 inches high, and is notched to clear the main chassis and the spacer bars and rotor arm of the bandswitch. The partition is held in place by a spade lug which passes through the chassis and by a mounting

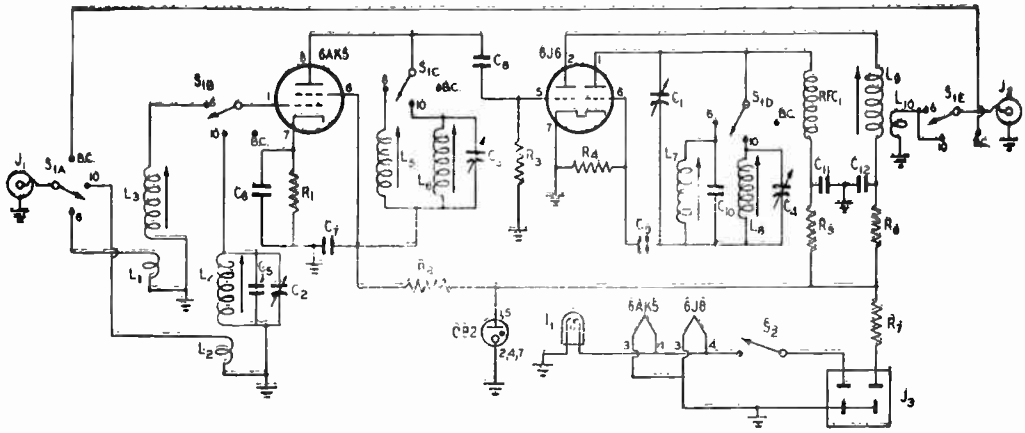


Fig. 12-23 — Circuit diagram of the bandswitching converter.

C_1 — 15-μμfd. variable reduced to one stator and 2 rotor plates (Millen 20015).
C_2, C_3, C_4 — 3–30-μμfd. mica trimmer (Millen 27030).
C_6, C_7 — 0.0015-μfd. ceramic (Centralab DA048002A).
C_8, C_9 — 100-μμfd. ceramic (Centralab CC32Z).
C_5, C_{10} — 10-μμfd. ceramic (Centralab CC20Z).
C_{11} — 500-μμfd. ceramic (Centralab D6501).
C_{12} — 0.01-μfd. ceramic (Centralab DA048003A).
R_1 — 220 ohms, ½ watt.
R_2, R_6 — 680 ohms, ½ watt.
R_3 — 1.5 megohms, ½ watt.
R_4 — 12,000 ohms, ½ watt.
R_5 — 47,000 ohms, ½ watt.
R_7 — 5000 ohms, 10 watts.
L_1, L_2 — 4 turns No. 28 d.s.c. close-wound over ground ends of L_3 and L_4.
L_3, L_4, L_5, L_6, L_7, L_8 — 6 turns No. 20 enameled wire close-wound on ⅜-inch diameter form; slug-tuned; inductance range 0.35 to 1.0 μh. (Cambridge Thermionic Corp. LS3 — 30 Mc.).
L_9 — Scramble-type winding on ⅜-inch slug-tuned form; inductance range 325 to 750 μh. (Cambridge Thermionic Corp. LS3 — 1 Mc.).
L_{10} — 20 turns No. 28 d.s.c. scramble-wound next to L_9.
I_1 — Adjustable-beam dial-light assembly.
J_1, J_2 — Coaxial-cable jacks (Amphenol 75-PC1M).
J_3 — 3-prong cable connector (Jones P-303AB).
RFC_1 — 300-μh. r.f. choke (Millen 34300).
$S_{1A, B, C, D, E}$ — 2-gang 6-circuit bandswitch (two Centralab SS sections).
S_2 — S.p.s.t. toggle switch.

lip which is screwed to the bottom side of the cabinet. It is located 3 inches in from the front edge of the chassis.

The heater switch and the pilot-light assembly are mounted at the lower left- and right-hand corners of the front panel with the bandswitch at the center, 1⅛ inches up from the bottom edge. The selector-switch index plate should have a rotor-shaft length of at least 3 inches, and the switch wafers should be mounted on the shaft with the first separated from the index plate by 1-inch spacers and with the second wafer separated from the first by 1⅝ inches.

The National MCN dial is centered above the bandswitch with the control shaft 3 inches above the bottom edge of the panel. It is wise to cut the large mounting hole suggested in the dial-mounting instruction sheet and then do the final fastening down of the dial after the tuning condenser and its mounting plate have been permanently secured in place.

The interior view of the completed converter shows the 6AK5 amplifier tube in front of the shield partition, with the grid inductances to

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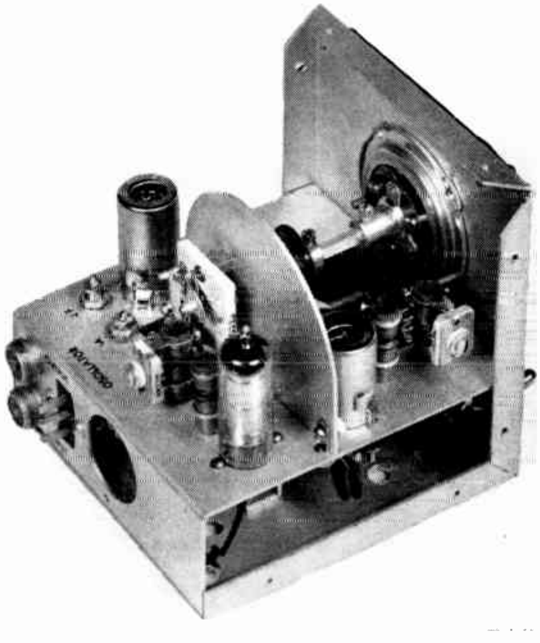


Fig. 12-24 — Interior view of the converter. Only the oscillator is tuned by the front-panel control, eliminating tracking problems.

the right of the tube. The padder condensers for 27 and 28 Mc. are mounted on the forward coil. From left to right across the rear of the chassis are the mixer-oscillator tube, five of the slug-tuned inductances, and the regulator tube. The i.f. output coil and the two oscillator coils are mounted below the chassis, as seen in the bottom view of the chassis subassembly. The r.f. plate coils are above the chassis to the left of the OB2 regulator, the 28-Mc. coil being the one with the trimmer condenser mounted across the terminals.

Construction will be simpler if the builder uses coils as shown. The Type LS3 30-Mc. inductors will resonate at 50 Mc. with the tube and circuit capacitances, and only a small padder capacitance is required to tune them to 27 and 28 Mc.

Coaxial jacks for the antenna and i.f. output cables are at the rear of the chassis to the left of the power-cable jack. They are closely grouped so that the input and output cables may be taped together to form a common cable.

Wiring can be done readily if the subassembly method is employed. The bottom view of the chassis, Fig. 12-25, shows how the circuit components are closely grouped around the tube sockets, with wiring completed to the point of making connections to the bandswitch. Twin-Lead of the 75-ohm type is used to make connection between the antenna input jack and the bandswitch. The two wires enclosed in spaghetti at the right of the chassis are the 6.3-volt leads which go to the heater switch.

Testing

The heater requirements of the converter are 6.3 volts at 0.625 amp., and the plate supply should deliver 200 to 250 volts at 25 to 30 ma. These may be drawn from the receiver with which the converter is to be used, or a separate supply may be employed. With power turned on, the plate voltage of the mixer and r.f. amplifier should measure 105 volts and the 6AK5 cathode resistor should provide a drop of approximately 2 volts. The 6AK5 cathode current should be about 8.5 ma. The regulator-tube drain will be about 8 ma.

Alignment of the converter is made most simple if a calibrated signal generator is available, otherwise amateur transmitter signals of known frequency may be used. The r.f. and i.f. circuits can be peaked on background noise. The oscillator stage should be on the low side of the signal frequency. It is possible to vary the bandspread of the converter over a wide range. With a fairly low order of padder capacitance, and with the inductance increased by the tuning slug, the 10- and 11-meter bands can be covered with one swing of the tuning dial. Anyone not interested in 11 meters can increase the bandspread on the 10-meter range by adding more padder capacitance and by decreasing the inductance of L_8. The converter as shown has 13 divisions of bandspread at 11 meters and 52 divisions at 10 meters, with the logging of frequencies made on the B scale of the dial. Bandspread for the 50-Mc. band is 48 divisions on the A scale. This spread may be increased by the same method.

Some operators favor a selected group of frequencies within a band. A slight improvement in the performance of the converter can be made in this case by peaking the r.f. amplifier circuits at a favorite spot rather than at the center of a band. There may be a tendency toward regeneration in the 50-Mc. r.f. amplifier, however, if the input and plate circuits are peaked at precisely the same frequency, making stagger tuning desirable.

Reducing Spurious Responses

In localities where there are stations operating in the high FM band a converter or receiver having broadband r.f. stages will experience considerable interference on the 50-Mc. range. This can be corrected in several ways, the simplest being the insertion of a 100-Mc. trap in the antenna lead, as described on page 389.

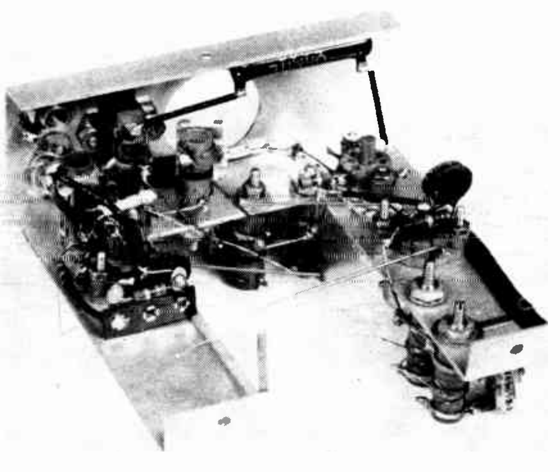


Fig. 12-25 — Construction of the converter is made easier if as much wiring as possible is done before the assembling is completed. This bottom view of the chassis subassembly shows the wiring completed to the point of connection to the bandswitch.

A 2-Meter Converter for Mobile Use

The converter shown in Figs. 12-26 through 12-29 was designed primarily for mobile reception in connection with a car broadcast receiver. It may also be used for home-station work, with any receiver capable of tuning to the high end of the broadcast band.

Because image rejection at 144 Mc. is low when an i.f. of 1600 kc. is used, double conversion must be employed for satisfactory results. Two 6J6 twin triodes are used, each as a mixer-oscillator. The first converts the signal frequency to 11.1 Mc., the second working from this frequency to 1600 kc. Only the high-frequency oscillator requires tuning during normal operation. Plate voltage for all circuits is stabilized by an 0B2 regulator tube.

Circuit Details

The first-mixer grid coil is tuned to the center of the 144-Mc. band by the tube and circuit capacitances. Its plate circuit is tuned to 11.1 Mc. by C_1 and L_3. The oscillator tunes from 132.9 to 136.9 Mc. to cover the band. The resulting i.f., 11.1 Mc., is then capacity-coupled by means of C_9 to the grid of the second mixer. C_4 is the bandset condenser and C_5 is the bandspread capacitor. No coupling condenser is needed between the oscillator and mixer.

The second 6J6 mixer-oscillator combination converts the 11.1-Mc. i.f. to 1600 kc. for working into a car radio. Note that a trap (C_2L_4) is connected in series with the coupling condenser between the two mixer circuits. This is tuned to 14.3 Mc. and attenuates image response at a frequency removed from the signal frequency by 3200 kc. This image, which falls within the 2-meter band when the converter is tuned to the low edge, can be reduced by 35 to 40 db. through adjustment of the trap.

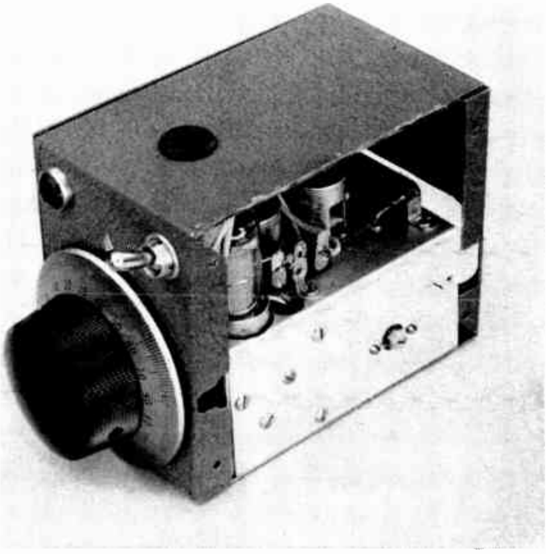
The plate circuit of the mixer is tuned to 1600 kc. by the trimmer, C_3, and a fixed capacitor, C_{11}. A low-impedance output link, L_6, terminates at J_2, and a short length of coaxial cable is used between the jack and the receiver.

Circuit details of the two oscillators are nearly identical, except that the low-frequency circuit uses only one capacitor, C_6, across the plate coil because the circuit operates at a fixed frequency of 12.7 Mc. Radiation from this oscillator, operated with 105 volts on to the plate, reached the high-frequency mixer and caused spurious responses. This condition was corrected by reducing the oscillator plate voltage by means of the dropping resistor, R_5, and by placing a copper shield between the two circuits. The reduction in oscillator signal made it necessary to add capacitive coupling between the oscillator and mixer. A 1½-inch length of 75-ohm Twin-Lead, identified as C_X on the circuit diagram, provides adequate coupling.

Construction

The chassis for the converter measures 1⅞ by 2⅞ by 4 inches and is made from a 6⅝ × 7¾-inch sheet of $\frac{1}{16}$-inch aluminum stock. A 1⅞-inch square is cut from each corner of the aluminum sheet so that the metal can be bent to form a boxlike chassis. It is recommended that the marking and drilling of mounting holes for parts be done before the chassis is bent into shape. The photographs of the converter show the location of most of the components. The hole for the oscillator bandset condenser (seen at the top of the chassis) is 1 inch square and is centered between the sides of the chassis.

Fig. 12-26 — A two-tube 144-Mc. converter for use with a car broadcast receiver. The side plate was removed to show the modification of the utility cabinet.



The mounting hole for the bandspread condenser is ¼ inch down on the front wall, and a ⅞-inch hole for the regulator-tube socket is centered to the left of the square hole. The high-frequency mixer-oscillator tube is centered on the chassis to the rear of the square hole, and the other r.f. tube is 1¼ inches to the right of the first tube. A mounting hole for the 11.1-Mc. coil is located ¾ inch in from the edge of the chassis directly to the right of the h.f. oscillator tube, and the 12.7-Mc. (second-oscillator) coil is ¾ inch in from the rear of the chassis and centered ⅞ inch away from the left edge. The form for L_5 is ⅝ inch from the right edge and ⅞ inch from the rear edge. R_8, J_1, J_2 and J_3 may be seen at the rear of the chassis and the location of these components is not critical. A two-terminal lug strip is located to the rear of the regulator tube for the leads running to the filament switch and the pilot-lamp socket. Trimmer condensers, C_1, C_3 and C_6, are mounted on the side walls of the chassis with their shafts 1⅛ inches from the top of the box. C_1, mounted on the left side, is ¾ inch back from the front wall and C_3 is 1⅜ inches farther toward the rear. C_6 is 1¼ inches from the rear wall on the right side. The mounting hole for the 14.3-Mc. coil is 7/16 inch up from the bottom edge of the chassis and is centered between C_1 and C_3.

The bottom view of the converter, Fig. 12-29, shows how the regulator-tube socket is mounted on a small aluminum bracket which is in turn mounted on the side wall of the chassis. An aluminum strip, 1 inch wide, should be bent to form a right angle and the position of the socket mounting hole should be marked after the bracket has been placed inside the chassis against the large clearance hole. Excess material may be cut from the bracket after it has been drilled for the socket. A three-terminal tie-point strip is mounted in a vertical position to the rear of the aluminum bracket, the bottom lug serving as a support point for the grid end of L_2. The coaxial cable and the antenna coupling loop are connected to the remaining two lugs.

The shield for the low-frequency oscillator circuit is made from a 1½ × 3¾-inch strip of 1/16-inch copper, bent to form a right angle having sides 1⅞ inches long and covering all of the components located at the top left-hand corner of the chassis. The shield is notched at the bottom corner to allow clearance for the coaxial cable which runs along the left edge of the chassis, and is equipped with a spade lug (the lug is soldered to the copper) for mounting.

The PRE-3 coil form for L_5 should be cut to 1¾ inches before the coil is wound. This and the other forms should then be marked and drilled to accommodate the windings. Terminal holes are drilled straight through the forms. A coat of cement may be applied to the windings and allowed to dry while other operations are performed.

As shown by Fig. 12-26, some work must be done on the metal utility box before it can be used as a cabinet. This consists of removing the top and bottom flanges at the right side of the case and notching the front and rear flanges to provide clearance for the condenser shaft and the jacks which are mounted on the aluminum chassis. A large slot must be cut in the rear of the case to allow access to the input and output jacks when the unit is assembled, and ¾-inch holes should be cut in the top, bottom and sides of the box so that the adjustment screws of the trimmer condensers may be reached with an alignment tool. The

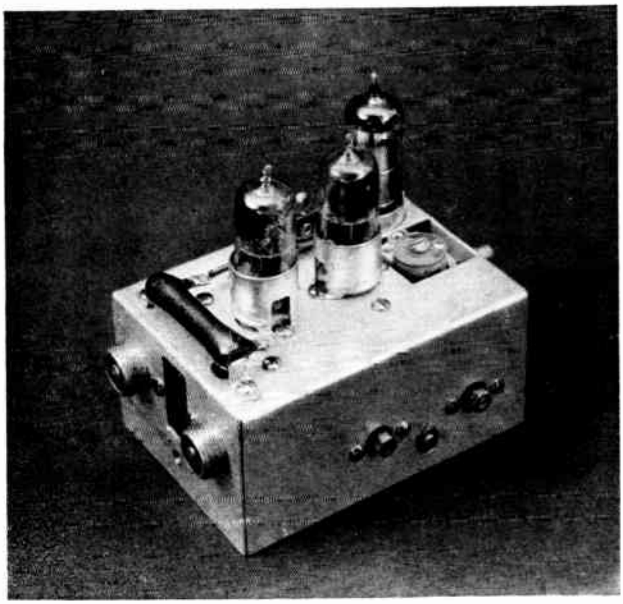


Fig. 12-27 — Top view of the 2-meter converter removed from its case.

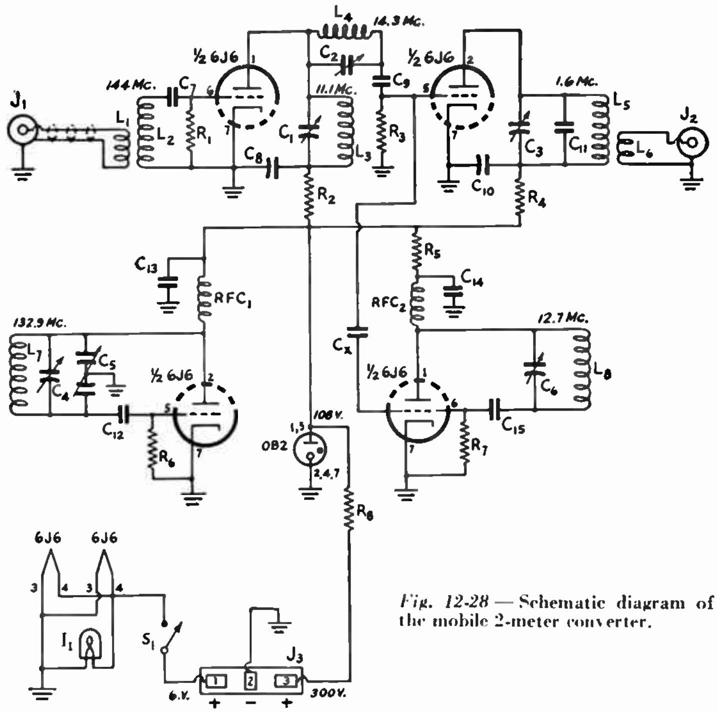


Fig. 12-28 — Schematic diagram of the mobile 2-meter converter.

C_1, C_3, C_6 — 62-$\mu\mu$fd. trimmer (Centralab 823-AZ).
C_2, C_4 — 20-$\mu\mu$fd. trimmer (Centralab 820-B).
C_5 — 5.27-$\mu\mu$fd. "butterfly" variable (Johnson 160-205).
C_7, C_{15} — 47-$\mu\mu$fd. mica.
C_8, C_{10} — 0.01-μfd. paper.
C_9 — 100-$\mu\mu$fd. mica.
C_{11} — 150-$\mu\mu$fd. mica.
C_{12} — 15-$\mu\mu$fd. mica.
C_{13} — 470-$\mu\mu$fd. mica.
C_{14} — 0.0047-μfd. mica.
C_X — Injection coupling, made from 75-ohm Twin-Lead — see text.
R_1, R_3 — 1.5 megohms, ½ watt.
R_2, R_4 — 1000 ohms, ½ watt.
R_5 — 0.22 megohm, ½ watt.
R_6, R_7 — 15,000 ohms, ½ watt.
R_8 — 3500 ohms, 10 watts.
L_1 — 4 turns No. 22 enam., close-wound, 3/16-inch diam.
L_2 — 6 turns No. 14 enam., 5/16-inch diam., ⅝ inch long.
L_3 — 20 turns No. 28 enam., ½-inch diam., 5/16 inch long. Coil wound on a National PRD-2 form.
L_4 — 28 turns No. 28 enam., ⅜-inch diam., ⅜ inch long. Coil wound on a National PRC-3 form.
L_5 — 75 turns No. 28 enam., 9/16-inch diam., 1 inch long. Coil wound on a National PRE-3 form.
L_6 — 10 turns No. 28 enam., close-wound over cold end of L_5.
L_7 — 3 turns No. 14 enam., 5/16-inch diam., approx. ½ inch long. See text for adjustment of length.
L_8 — 20 turns No. 28 enam., ½-inch diam., 5/16 inch long. Coil wound on a National PRD-2 form.
I_1 — 6.3-volt pilot-lamp assembly.
J_1, J_2 — Coaxial-cable jack (Amphenol 75-PC1M).
J_3 — Three-prong cable jack (Jones S-303-AB).
RFC_1 — 1-μh. r.f. choke (National R-33).
RFC_2 — 300-μh. r.f. choke (Millen 34300).
S_1 — S.p.s.t. toggle switch.

heater switch and the pilot lamp are mounted as far toward the top of the front panel as possible, and a ¾-inch hole is drilled up from the bottom of the panel for a distance of 1½ inches. This large hole will allow the National AM dial to be positioned correctly with respect to the tuning-condenser shaft after the chassis has been placed inside the cabinet.

The miniature Johnson condenser, C_5, may have a small-diameter control shaft that does not fit a standard dial coupling, in which case a bushing or shim is required. Fortunately, a ¼-inch length of ¼-inch soft copper tubing can be made to fit the shaft by working the inner surface with a rattail file.

Wiring

Construction and wiring are not difficult if the parts are mounted and wired in the following order: First, mount the tube sockets, the three jacks, and the lug strip (the one located on the top of the chassis). Next, complete the heater wiring and mount the grid-leak resistors in place. C_4 can now be soldered across the terminals of C_5 and L_7 can also be mounted on the condenser. This assembly is then mounted on the front wall of the chassis and, in turn, is connected to the tube socket by means of a short length of stiff tinned wire at the plate side and by C_{12} at the grid side. Now, mount the vertically-positioned lug strip on the side wall and connect a short piece of coaxial cable between the top lugs and J_1. C_7 can now be connected between the tube socket and the terminal strip and L_2 (with the small antenna winding slipped inside the cold end of the coil) may be mounted.

Condensers C_1, C_3 and C_6, and coils L_3, L_5

and L_8, are now mounted and wired into their respective circuits and, from here on, the wiring can proceed in any order. The 0.01-μfd. by-pass condensers are mounted in a vertical plane next to C_1 and C_3, respectively, and RFC_1 and R_2 are supported at the B-plus end by Pin 5 of the regulator-tube socket. The small metal post at the center of the rear tube socket is used as the tie point for the common connection between C_{14}, R_5, RFC_2 and the plate-voltage lead. L_4 is wired to C_2 after the padder condenser has been mounted between the coupling condenser, C_9, and a piece of No. 12 tinned wire which runs down to the stator terminal of C_1.

If the constructor wishes to use noise as a means of making a rough alignment of the converter, it is suggested that the injection-voltage condenser, C_X, and the dropping resistor, R_5, be left out of the circuit at this time. Of course, the plate of the 6J6 must be connected directly to RFC_2 in this case. The converter will have a much higher noise level when wired in this manner and alignment on noise is simplified. Actually, this is a poor method of aligning a double converter and should be used only as a last resort.

Testing

Power requirements for the converter are approximately 300 volts at 50 ma. and 6 volts at 0.9 ampere. A receiver capable of tuning to 1600 kc. should be coupled to the converter by a short length of coaxial cable and the receiver adjusted for normal operation at this frequency. If a signal generator is to be used, it is connected to the input jack, J_1, and if a generator is not available, the converter should be coupled to a low-impedance antenna system. Remember that C_X and R_5 should both be incorporated in the circuit if the converter is to be aligned with the aid of a test signal.

If preliminary testing is to be done with noise, the converter and the receiver are turned on and the converter output tuning condenser, C_3, adjusted until the noise level is at maximum. The low-frequency oscillator should now be adjusted by means of C_6 until a further increase in noise level is heard. C_4, the h.f. oscillator padder, should also be adjusted to produce maximum receiver output and this should occur with the padder adjusted to approximately half capacity.

At this point, it is necessary to introduce a test signal of known frequency, and it is helpful if the signal can be set at 146 Mc. — the center of the band. With C_5 set at half capacity, C_4 is adjusted until the test signal is heard. It is advisable to check the frequency of the high-frequency oscillator at this point to make sure that it is adjusted to the low-frequency side of the input mixer circuit. Condensers C_1, C_3 and C_6 should now be tuned for maximum converter sensitivity. The frequency of the second oscillator can be checked by tuning the range around 12.7 Mc. with an all-band receiver.

The converter bandspread can be adjusted by changing the L/C ratio of the first oscillator, by altering the spacing between turns of L_7. Of course, C_4 must be reset each time the inductance of the coil is varied. Because the first mixer has a broad frequency response, it is only necessary to peak the input coil, L_2, at the center of the band by varying the length of the coil. The coupling between the antenna link and L_2 should be adjusted for maximum response.

When all of the circuits have been aligned, it is time to adjust the 11.3-Mc. trap. With a strong test signal at the low end, tune to the high side until the image is heard, and then adjust C_2 for lowest image response.

It is to be expected that the various circuits will need slight readjustment after the chassis has been enclosed in the cabinet. However, this presents no difficulty as all of the tuning controls are accessible through small holes in the cabinet walls.

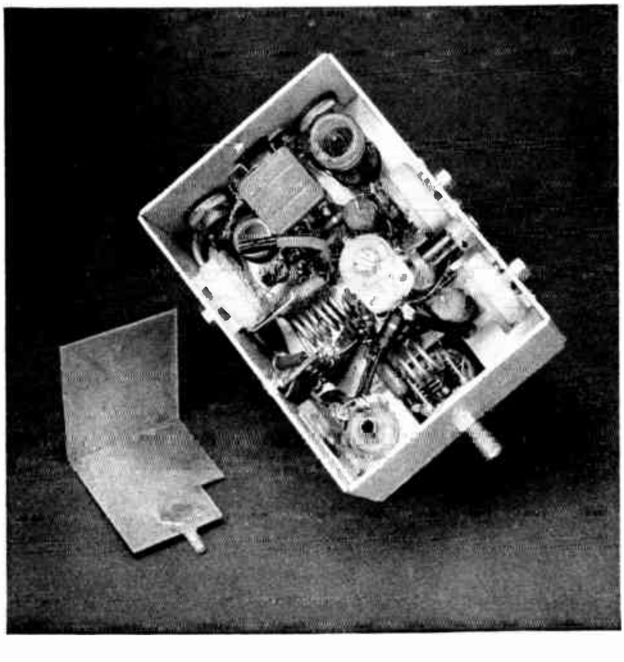


Fig. 12-29 — Bottom view of the 2-meter converter, showing the small copper shield used to reduce "birdies" from the low-frequency oscillator.

A 6J6 Preamplifier for 28, 50 and 144 Mc.

The triode preamplifier shown in Figs. 12-30 to 12-33 will improve the sensitivity and lower the noise figure of receivers and converters that are deficient in these characteristics. It uses a 6J6 as a push-pull neutralized amplifier, with plug-in coils in its grid and plate circuits. A self-contained power supply is included, so the only connections needed are to the receiver antenna terminals and the a.c. line.

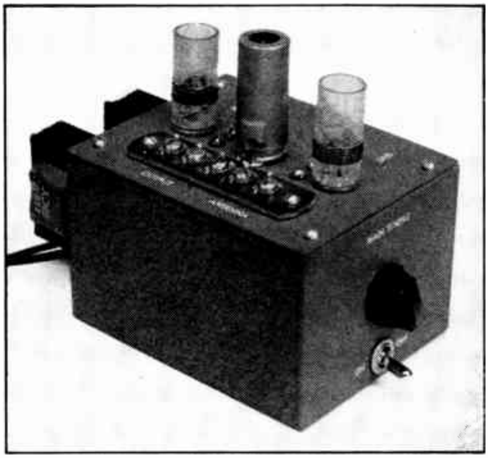


Fig. 12-30 — An r.f. preamplifier for 28, 50 and 144 Mc. The 50-Mc. coils are shown.

The r.f. components are mounted on the top plate of a standard utility box, 3 by 4 by 5 inches in size. The power-supply parts are attached to the walls of the box itself. The 6J6 socket is in the middle of the top plate, with the plug-in coil sockets equally spaced in front and back of it. The butterfly tuning condensers are on the underside of the same plate, as close as possible to the coil sockets. The neutralizing trimmers mount directly on the stators of the tuning condensers.

The power supply uses two small 6.3-volt filament transformers wired "back-to-back," a selenium rectifier, two small filter condensers, and a resistor in lieu of a choke. The filament transformers also supply the heater voltage for the 6J6. Fig. 12-33 shows the utility box with all power-supply components mounted in place and wired, ready for use.

Adjustments

The amplifier must be neutralized before operation can be checked. This may be done in two ways. The neutralizing trimmers should be set near minimum capacitance and the tuning-condenser gang turned through its entire travel, while listening on the receiver with which the amplifier is to be used. The output terminals of the amplifier should be connected to the antenna terminals of the receiver by a short length of 300-ohm line, and an antenna of the type normally used for the band in question should be attached to the preamplifier. If no antenna is available a carbon resistor of the value of the line impedance (75, 300, 500 ohms, etc.) should be connected across the amplifier input terminals. Moving the neutralizing trimmers either way from the proper setting will cause the 6J6 to oscillate, as indicated by excessive noises in the receiver. Best operation will be had with the trimmers at the mid-

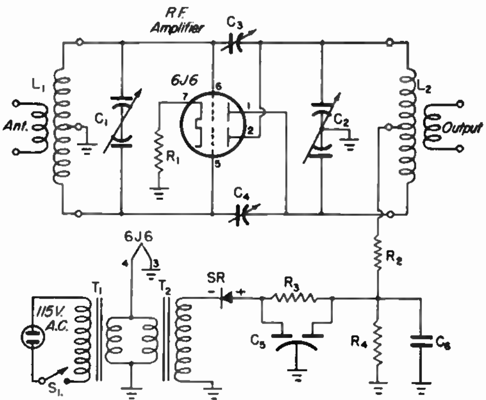


Fig. 12-31 — Schematic diagram of the 3-band r.f. preamplifier.

C_1, C_2 — 15-μμfd. butterfly-type variable (Hammarlund BFC-12). Flexible coupling is National type TX-10.
C_3, C_4 — 3–30-μμfd. mica trimmer.
C_5 — 40/40-μfd. 150-volt electrolytic.
C_6 — 100-μμfd. mica.
R_1 — 47 ohms.
R_2 — 220 ohms.
R_3 — 1000 ohms.
R_4 — 0.1 megohm.
S_1 — S.p.s.t. toggle.
SR — Selenium rectifier (Federal 402D3150-A).
T_1, T_2 — 6.3-volt 1-amp. filament transformer (Merit P-2944).

COIL DATA FOR THE 6J6 PREAMPLIFIER

Band	*Antenna*	*Grid, L_1*	*Plate, L_2*	*Output*
28 Mc.	3 t. No. 18 e. ⅜-inch dia. inside L_1.	14 t. No. 24 e., c.t., ⅝ inch long.	Same as L_1.	6 t. No. 18 e. ⅜-inch dia. inside L_2.
50 Mc.	4 t. No. 18 e. 5/16-inch dia. inside L_1.	6 t. No. 24 e., c.t., 5/16 inch long.	Same as L_1.	6 t. No. 18 e. 5/16-inch dia. inside L_2.
144 Mc.	2 t. No. 18 e. ¼-inch dia. Insert between sections of L_1.	2 t. No. 16 t. each side of c.t., 5/16-inch dia., ⅝ inch long.	Same as L_1, but ⅜ inch long.	3 t. No. 18 e. ¼-inch dia. Insert between sections of L_2.

Coil forms are ⅝-inch diameter, 5 prong (Amphenol 24-5H) with sockets to match (Amphenol 54-5H). The 144-Mc. coils are air-wound, using cut-down forms for bases.

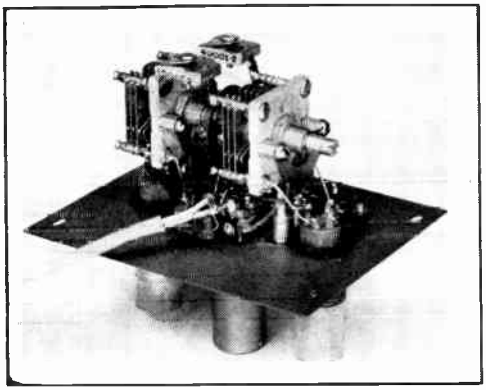


Fig. 12-32 — The r.f. portion of the 3-band preamplifier is mounted on the cover plate of the utility box.

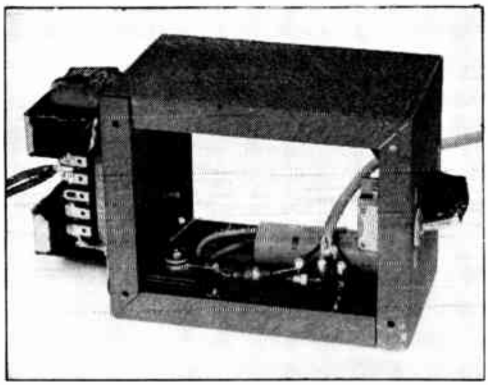


Fig. 12-33 — Power-supply components of the preamplifier are mounted on the walls of the utility box.

point between the settings at which oscillation starts. If the normal minimum capacitance of the trimmers is too high to permit neutralization the movable plates should be cut down in size.

The most effective check for neutralization is had by inserting a burned-out 6J6 (or one with a heater prong cut off) in the socket and adjusting the trimmers for *minimum* response while listening to a strong signal. With some care it is possible to find a setting that holds for all three bands, but the adjustment should be made for the band on which best weak-signal reception is desired.

No provision is made for padding the coils, so the inductance should be close to the correct value. This may be checked by inserting an iron core into the plate and grid coils, one at a time. If an increase in signal results the inductance of the coil in question is too low. As various antennas and receiver input circuits may reflect different loads back on L_1 and L_2 this check should be made with the receiver and antenna with which the amplifier is to be used.

The coil and condenser values given represent a compromise for three-band operation. If such a preamplifier is to be used for 144 Mc. only improved results can be achieved by using variable condensers of lower minimum capacitance and eliminating plug-in coils. The reduced circuit capacitance thus obtained will permit the use of more efficient coils for the 144-Mc. band.

CHAPTER 13

V.H.F. Transmitters

Beginning with the v.h.f. region, amateur frequency assignments are not in direct harmonic relationship with our lower-frequency bands. This fact, coupled with the necessity for extreme care in selection and placement of components for low circuit capacitance and minimum lead inductance, makes it highly desirable to construct separate gear for v.h.f. work, rather than attempt to adapt for v.h.f. use a transmitter designed for the lower amateur frequencies.

Transmitter stability requirements for the 50-Mc. band are the same as for lower bands, and proper design may make it possible to use the same rig for 50, 28, 21, and even 14 Mc., but incorporation of 50 Mc. and higher in the usual multiband transmitter is generally not feasible. Rather, it is usually more satisfactory to combine 50 and 144 Mc., since the two bands are close to a third-harmonic relationship. At least the exciter portion of the transmitter may be made to cover the requirements for both these bands very readily.

Though no stability restrictions are imposed by law on operation at 144 Mc. and higher amateur bands (other than that the entire emission must be kept within the limits of the band in question), experience has demonstrated the value of using crystal control or its equivalent for at least home-station operation in the 2-meter band. When large numbers of stations flock to a v.h.f. band, as occurred in the first months of operation on 144 Mc., severe interference soon develops if unstable transmitters and broadband receivers are employed. Conversion of this activity to crystal-controlled transmitters and receivers having the minimum bandwidth necessary for voice communication makes it possible for hundreds of stations to operate without undue interference, in the same band which appeared over-crowded with only a dozen or so stations working with inferior gear.

The use of narrow-band communications systems also pays off in the form of improved efficiency in both transmitter and receiver. It is this factor, perhaps more than the interference potentialities of the wide-band systems, which makes it desirable to employ advanced techniques at 220 and even 420 Mc. Stabilized transmitters for 220 Mc. are not too difficult to build, and their use at this frequency is highly recommended.

Construction of multistage rigs for 420 Mc. is not easy, and the choice of tubes suitable for this type of work is quite limited, but the advanced amateur who is interested in making the most of the interesting possibilities afforded by this developing field will be satisfied with nothing less. The 420-Mc. band is much wider than our lower v.h.f. assignments, however, and interference is not likely to become a limiting factor in this band for a long time to come. Thus it may be more important, in many localities, to get activity rolling with any sort of gear, leaving perfection in design to come along as the need develops.

At 420 Mc. and in the higher amateur assignments most standard tubes cannot be used with any degree of success, and special tubes designed for these frequencies must be employed. These types have extremely-close electrode spacing, to reduce transit-time effects, and are constructed with leads having virtually no inductance. Several more-or-less conventional tubes are now available which will operate with fair efficiency up to about 500 Mc., and the disk-seal or "lighthouse" variety will function up to about 3000 Mc. with specially constructed circuits. Above about 2000 Mc. the most useful vacuum tubes are the klystron and the magnetron. These are essentially one-band devices, the frequency-determining circuits being an integral part of the tube itself. Tuning over a small frequency range, such as an amateur band, is possible, usually by warping a built-in cavity, but the tubes are not independent of frequency in the conventional sense.

Frequency modulation may be used throughout the v.h.f. and higher bands, wide-band emission being permitted above 52.5 Mc. and narrow-band FM above 51 Mc. Where suitable receivers are available to make best use of such emissions, either wide-band or narrow-band FM can provide effective v.h.f. communication, and the latter is becoming increasingly popular, particularly in congested areas, where its freedom from broadcast interference permits operation under conditions which would be prohibitive for amplitude-modulated transmitters of any appreciable power.

In areas where there is television service in operation, the v.h.f. enthusiast must guard against interference to television reception. One way of keeping TVI to a minimum is the use of low power in the driving stages, building up the power level only after the operating frequency is reached. Extensive shielding and filtering, covered in other chapters, may also be required.

A 400-Watt Transmitter for 50 and 144 Mc.

A high-powered transmitter for use on our two most-popular v.h.f. bands presents some knotty design problems. It is not always easy to develop satisfactory drive for the higher band, and an efficient band-changing system for a 144-Mc. amplifier calls for something better than the ordinary plug-in coil arrangement. These two factors were prime considerations when the all-tetrode rig for 50 and 144 Mc. shown in Figs. 13-1 to 13-7 was laid out.

The exciter has separate output stages for the two bands, eliminating the necessity for driving the final stage with a frequency multiplier on the higher one. Efficient operation of the final stage is attained with a novel form of tank circuit that avoids the use of a plug-in coil for 144 Mc. As a result, the transmitter has practically the same over-all efficiency as would be obtainable if it were designed for either band alone.

THE EXCITER

Though the two units were intended for use together as a complete 400-watt transmitter, as shown in the composite photograph, the exciter portion may be used as a low-powered transmitter by itself. As an exciter it has the virtue of providing uniform drive for the final on both bands. Other points of interest include quick band changing, crystal switching, VFO-input provision, low power consumption, and freedom from critical adjustments.

The circuit diagram of the exciter is given in Fig. 13-3. The 6AR5 Tri-tet oscillator employs a fixed-tuned cathode circuit, C_8L_3. The plate circuit, C_1L_1, tunes 24 to 27 Mc., the oscillator tripling when 8-Mc. crystals are used and quadrupling with 6-Mc. crystals. Five crystals are provided for by the switching circuit, and a sixth position of the switch connects the 6AR5 grid to a tuned circuit, C_5L_1, which is in turn link coupled to the VFO-input jack, J_1. Switch S_{2a} grounds the cathode of the oscillator tube when VFO input is used. The second 6AR5 is a frequency doubler with its output link coupled to an 832A amplifier-tripler circuit.

As a straight-through amplifier at 50 Mc., the 832A uses a low-value grid resistor, R_5, cut into the circuit by switch S_{3a}. A high-resistance grid-leak, R_6, is picked up by S_{3a} when the tube is operated as a frequency tripler to 144 Mc. Tube and circuit capacitance resonate the grid coil, L_8, at approximately 49 Mc. Jacks J_2 and J_3 permit metering of the grid and the cathode currents with J_3 also serving as the keying jack for c.w. work at 50 Mc. The plate circuit uses plug-in coils with the output link-coupled to the final by means of L_{10} in the 50 Mc. coil. At 144 Mc., output is capacity-coupled to the 2-meter output stage by condensers C_{15} and C_{16}. The 144-Mc. stage, also an 832A, has grid and cathode jacks as in the previous stage. It is made active by applying heater voltage through S_{3b}.

Power wiring for the unit is shown in the lower section of Fig. 13-3. Power for the exciter is fed through a 5-prong male receptacle. A 4-prong female receptacle permits taking out heater and plate voltages for an external VFO. Changing from VFO to

Fig. 13-1 — A complete 400-watt transmitter for 50 and 144 Mc.

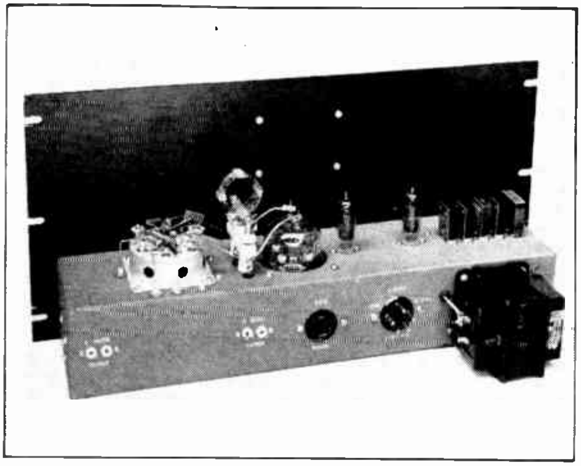


Fig. 13-2 — A rear view of the 50- and 144-Mc. exciter. Across the top of the chassis, from right to left, are the crystal sockets, the oscillator and doubler tubes, the 832A amplifier-tripler and its plate coil, and the inverted 144-Mc. amplifier stage. Crystal sockets, used as r.f. output terminals, are mounted on the rear wall of the chassis along with the power plugs and the filament transformer.

crystal operation is done by means of the crystal switch and $S_{2a, b}$.

Higher plate voltage is applied to the 144-Mc. amplifier than is used with the other three circuits, making the output on 144 Mc. comparable with that of the 50-Mc. amplifier.

Construction

The exciter is built on a metal chassis measuring 3 by 5 by 17 inches. The aluminum rack panel, ⅛ by 8¾ by 19 inches in size, is held in place by the mounting nuts of the various controls.

Plate tuning condensers for the oscillator and the doubler are mounted on the front wall of the chassis. These two controls are hot with +300 volts and must be insulated from the chassis. Bakelite tuning knobs without metal dial plates protect the operator.

The amplifier-tripler circuit, located at the left-center of the chassis as seen from the rear view, has its plate coil mounted on a National type XB-16 socket. Shield braid is used for the connections between the coil socket and the 832A plate caps, while Twin-Lead is wired between the output link and the output terminals. The tube is submounted on a Johnson shielded socket, Type 122-101, and the plate tuning condenser, C_3, is mounted to the left of the tube socket on an aluminum bracket.

The 144-Mc. amplifier has the shielded tube socket mounted in an inverted position. The grid chokes, RFC_5 and RFC_6, are mounted between the socket terminals and a tie-point strip which is in turn mounted on the metal part of the socket along with the button-type by-pass condensers. The coupling condensers, C_{15} and C_{16}, are between the tube socket and the amplifier-tripler plate coil socket. Millen No. 32150 through-bushings, set in the chassis to the left and rear of the tube socket, pass d.c. and heater leads for the 832A.

The bottom view of the exciter shows the plate tuning condenser, C_1, mounted on the end wall of the chassis just below the two-terminal tie-point strip which supports the output link, L_{12}. A heavy copper strip is used as the ground lead for the rotor of the tuning condenser. The screen-dropping resistor is mounted on a tie-point strip located on the rear wall of the chassis.

Testing

Power-supply requirements for the exciter will depend on how the unit is operated. If it is to serve as a low-power transmitter, the supply need deliver only 300 volts at approximately 175 ma. For exciter service, two supplies are recommended — one delivering 300 volts at 125 ma. and one furnishing 400 volts at 100 ma., the latter to be used on the second 832A. The filament transformer must deliver 6.3 volts at 4 amp. in either case.

If operation with a VFO not having its own supply is contemplated, the power-supply capabilities should be increased to meet the extra requirements. When the V.H.F. Man's VFO, Figs. 13-8 to 13-10, is used it increases the heater load by 2 amp. and the plate-current drain by approximately 60 ma.

Performance of the oscillator and the doubler circuits should be checked first. This is done with the plate and screen voltages removed from both 832 stages, and with a low range milliammeter plugged in J_2. The oscillator cathode switch should be opened. Table 13-I will assist in the selection of a crystal for the desired output frequency, and shows the

TABLE 13-I

Crystal	*Oscillator*	*Doubler*	*Amplifier-Tripler*	*Amplifier*
6250	25	50	50	
6750	27	54	54	
8333.4	25	50	50	
9000	27	54	54	
6000	24	48	144	144
6166.6	24.6	49.3	148	148
8000	24	48	144	144
8222.2	24.6	49.3	148	148

NOTE: Crystal frequencies in kc.; other frequencies in Mc.

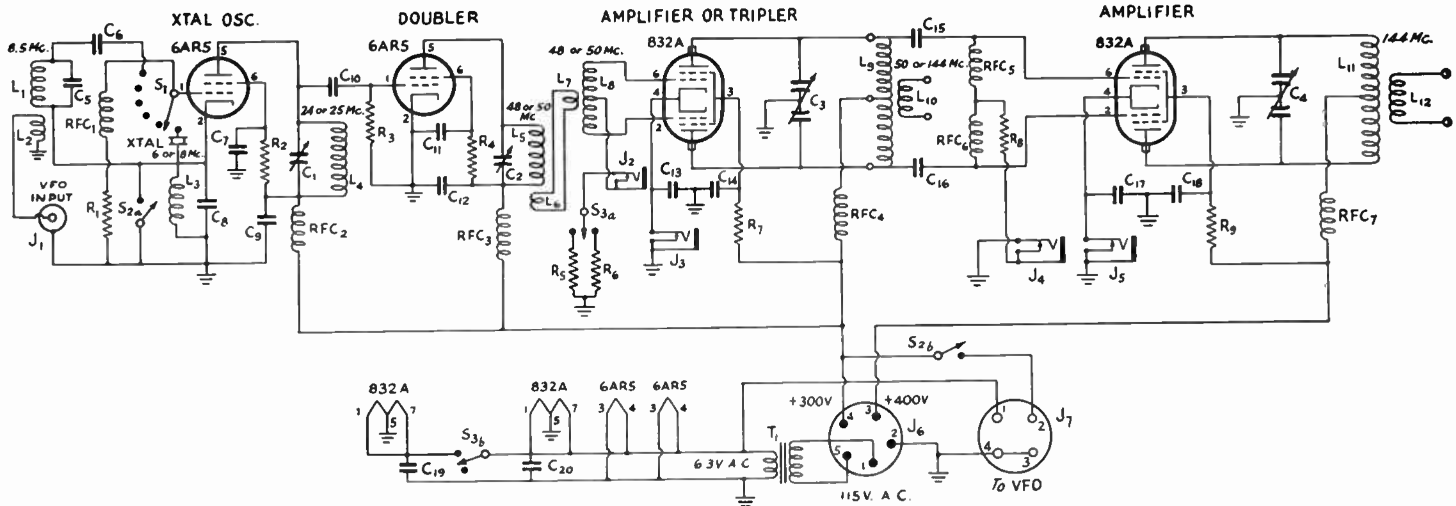


Fig. 13-3 Circuit diagram of the 50-144 Mc. exciter.

C_1, C_2 — 25-$\mu\mu$fd. variable (Millen 20025).
C_3, C_4 — 25-$\mu\mu$fd.-per-section split stator (Bud LC 1661).
C_5 — 22-$\mu\mu$fd. midget mica.
C_6, C_{10} — 100-$\mu\mu$fd. midget mica.
C_7, C_9, C_{12} — 0.0047-μfd. mica.
C_8 — 68-$\mu\mu$fd. mica.
C_{11}, C_{13}, C_{14}, C_{20} — 470-$\mu\mu$fd. midget mica.
C_{15}, C_{16} — 10-$\mu\mu$fd. midget mica.
C_{17}, C_{18}, C_{19} — 500-$\mu\mu$fd. button-type by-pass.
R_1 — 0.12 megohm, ½ watt.
R_2 — 15,000 ohms, 1 watt.
R_3 — 47,000 ohms, ½ watt.
R_4 — 22,000 ohms, 1 watt.
R_5, R_8 — 22,000 ohms, ½ watt.
R_6 — 0.1 megohm, ½ watt.
R_7, R_9 — 25,000 ohms, 10 watts.
L_1 — 18 turns No. 24 enam., 3/8 inch long, 1-inch diam.
L_2 — 4 turns No. 24 enam., close-wound at ground end of L_1.
L_3 — 14 turns No. 20 tinned, 7/8 inch long, 5/8-inch diam.
L_4 — 10 turns No. 20 tinned, 5/8 inch long, 5/8-inch diam.
L_5 — 5 turns No. 20 tinned, 5/16 inch long, 5/8-inch diam.
Note: B & W Miniductor No. 3007 used for L_3, L_4 and L_5.
L_6, L_7 — Two-turn coupling links.
L_8 — 18 turns, No. 20 enam., 5/8 inch long, ½-inch diam.
L_9 — 50 Mc.: 4 turns No. 20 enam., ¾ inch long, 1¼-inch diam. National type AR-16-10C with 2 turns removed from each end.
— 144 Mc.: 4 turns No. 14 tinned, 7/8 inch long, ¼-inch diam.
L_{10} — 50-Mc. output link: 2 turns No. 20 enam., wound around L_9.
L_{11} — 4 turns No. 12 tinned, 5/8-inch diam., wound in two sections with two turns each side of center tap and a 3/8-inch space at center, turns spaced wire diam.
L_{12} — 144-Mc. output link: 2 turns No. 14 tinned, ½-inch diam., turns spaced wire diam.
J_1 — Coaxial-cable connector.
J_2, J_3, J_4, J_5 — Closed-circuit jacks.
J_6 — 5-prong male receptacle.
J_7 — 4-prong female receptacle.
RFC_1 — 2.5-mh. r.f. choke.
RFC_2, RFC_3, RFC_4 — 7-μh. r.f. choke (Ohmite Z-50).
RFC_5, RFC_6, RFC_7 — 1.8-μh. r.f. choke (Ohmite Z-144).
S_1 — 8-position selector switch (Amphenol 36-1).
S_{2a}, S_{2b} — D.p.s.t. toggle switch.
S_{3a}, S_{3b} — D.p.d.t. toggle switch.
T_1 — Filament transformer: 6.3 volts, 6 amp.; see text.

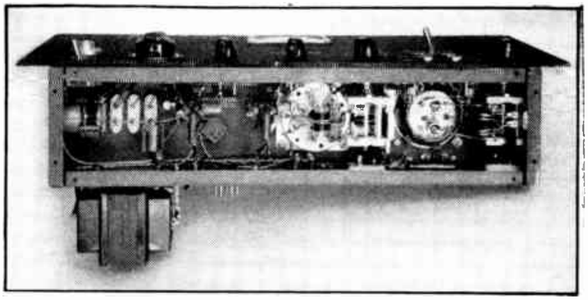


Fig. 13-4 — Bottom view of the v.h.f. exciter. The VFO input coil is at the left end of the chassis. Plate coils for the oscillator, the doubler and the 144-Mc. amplifier circuits are mounted on the tuning condensers. The grid coil for the amplifier-tripler stage is mounted on the tube-socket terminals.

frequencies to which the various circuits should be tuned. With plate voltage applied and with the doubler tuned to resonance, the grid current of the 832A should be approximately 7 ma. when an 8-Mc. crystal is used. Grid current will be 5 or 6 ma. with a 6-Mc. crystal. Total cathode current for the two 6AR5s should be 50 ma. Normal screen voltage for the oscillator and the doubler tubes is about 230 and 200 volts, respectively.

The 832A may now be tested at 50 Mc. This requires a 100-ma. meter in the cathode circuit and a 10-watt lamp coupled to the output terminals. When the plate circuit is tuned to resonance, the grid current should stay up around 5 ma., the cathode current should dip to about 65 ma., and the lamp should indicate an output of 6 to 8 watts. A screen potential of 160 volts is correct with the amplifier loaded. The plate current should rise noticeably and the grid current fall to zero when excitation is removed. This last test must be one of short duration.

To check the 144-Mc. stage, plug in the 2-meter coil at L_{11} and apply the heater voltage through S_{3b}. Grid current for the amplifier will be around 3.5 ma. A recheck of the tripler should show a grid current of 1 ma. and a cathode current of 55 to 60 ma.

With a 400-volt supply connected to the amplifier and with the dummy load across the 144-Mc. output terminals, 6 to 8 watts output should be obtained with an 832A cathode current of approximately 65 ma. Grid current should be 3 ma. and the screen voltage should measure 170 volts. A short test for self-oscillation should be made by removing the excitation.

The general method of tuning does not change when a VFO is used as the frequency-control unit. However, it is important that the oscillator cathode switch be closed; otherwise the oscillator circuit will take off on its own.

It is recommended that a calibrated wavemeter be used to check the tuning adjustments, particularly those associated with 144-Mc. operation. There are numerous out-of-band harmonics from the low-frequency crystals and the high order of frequency multiplication. Be careful to choose the proper harmonics in the first two stages.

THE POWER AMPLIFIER

Customary plug-in coil arrangements are not well adapted to use in high-power 144-Mc. stages. The lead inductance and parallel capacitance inherent in the best jack bars and coil bases leave almost nothing for the coil itself, with the result that efficient operation is all but impossible. The 144-Mc. tank circuit used here is, however, practically as effective as if it were designed for one-band operation. When the amplifier is used on 144 Mc. the plate circuit operates as a conventional tuned quarter-wave line. In changing to 50 Mc. it is merely necessary to remove the shorting bar, change the grid coil, and plug the 50-Mc.

Fig. 13-5 — Rear view of the 4-65A amplifier, showing the two-band tank circuit set up for 50-Mc. operation. R.f. input terminals are on the rear wall to the left and receptacles for the power leads are to the right. The 144-Mc. output terminals are on a bracket to the left of the protective tube. The 50-Mc. output terminal is mounted directly on the XB-15 socket for the plate coil. A plug-in shorting bar, used across the plate lines at 144 Mc., is shown in the foreground.

A V.H.F. Man's VFO

Though a VFO is considered to be an almost indispensable part of an amateur station for lower frequencies, v.h.f. operation is still carried on mainly with crystal control. This is largely because of the relatively lower occupancy of the v.h.f. bands and the freedom from interference problems which results. It is also, in part, the result of the fact that, as we go higher in frequency, it becomes more difficult to generate an entirely satisfactory signal by means other than with crystal control.

With proper attention to the well-known factors affecting oscillator stability a frequency control unit for 80-, 40- or 20-meter use can be built with a minimum of complications, but many a signal which sounds acceptable on these frequencies becomes quite fuzzy by the time it is multiplied to the v.h.f. bands. Even on 10 meters it is not too easy to obtain a pure d.c. note, especially when the oscillator frequency is modulated for narrow-band FM.

The frequency-control unit described herewith has a degree of frequency stability that is adequate for the high-order frequency multiplication required in v.h.f. service, and the design of the audio portion is such that little or no hum is introduced in the reactance-modulation process. The unit has the reactance-modulator and speech amplifier built in, the gain of the latter being only just enough to provide sufficient deviation for 10-meter NFM. Much of the hum present on some FM signals comes from the use of excessive speech gain, or haywire patching systems in order to utilize the speech equipment in some other portion of the transmitter.

This unit, shown in Figs. 13-8–13-10, was designed with the needs of the v.h.f. man in mind. Since many v.h.f. operators also work on 10 and 11 meters the oscillator tuning range was extended to include these bands, as well as 2 and 6 meters. The actual output frequency of the VFO is 6.74 to 9 Mc. It is designed to serve as a crystal substitute, and may be plugged into the crystal socket of any transmitter employing crystals falling within its tuning range. Thus, though the dial is calibrated only for the bands from 11 to 2 meters, the unit may be used on 40 or 20, or on portions of the higher v.h.f. bands that are in harmonic relationship with the output frequency. The output is sufficient so that the unit may also be used as a driver for a low-powered amplifier or frequency multiplier whose grid circuit is on that frequency. It also includes a reactance modulator and speech amplifier, providing narrow-band FM on 27 Mc. and higher frequencies with only the addition of a crystal microphone.

Two 6AG7s are used in the r.f. portion. The first is an oscillator-doubler employing the highly-stable Clapp oscillator, the operating frequency of which is 3370 to 4500 kc., doubling in the plate circuit. The second is an amplifier operating on 6.74 to 9 Mc. By means of separate padders switched in by a front-panel control, a reasonable amount of bandspread is provided for each of the four bands from 2 to 11 meters. The 50-Mc. band covers 55 divisions on the vernier dial, 144 Mc. is covered in 25 divisions, the 10-meter band occupies 80 divisions, and 11 meters 20 divisions. By proper setting of the padders the 2- and 11-meter ranges can be made to come at the opposite ends of the National MCN dial, leaving the two other spaces on the dial card for the 10- and 6-meter calibrations.

Frequency modulation is accomplished by means of a reactance modulator and a speech amplifier, both using 6BA6 miniature tubes. Deviation of the oscillator frequency is approximately 500 cycles, providing adequate swing for 10-meter NFM as a result of the eight times multiplication. A deviation of approximately 10 kc. is possible in the 6-meter band, and as much as 30 kc. on 2 meters. This greater

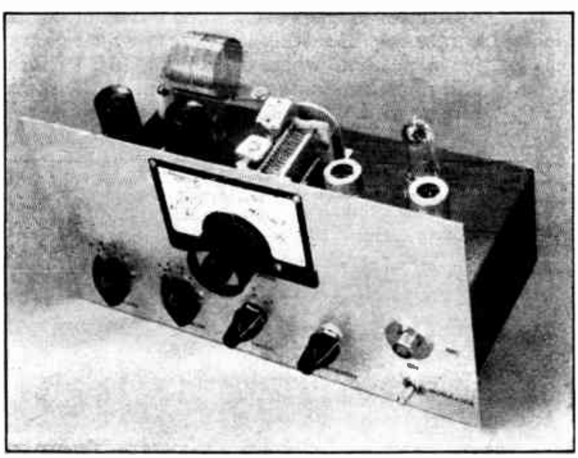


Fig. 13-8 — Panel view of the v.h.f. VFO with NFM modulator.

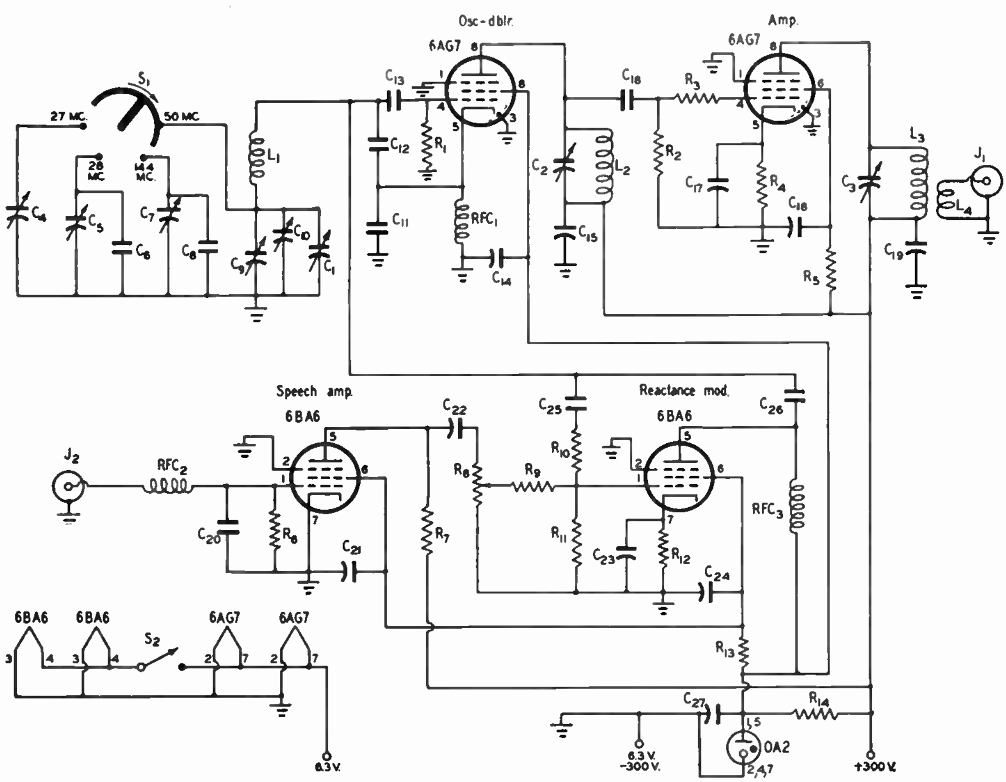


Fig. 13-9 — Circuit diagram of the NFM control unit for v.h.f. use.

C_1 — 35-$\mu\mu$fd. variable, double spaced (Millen 21935).
C_2, C_3 — 100-$\mu\mu$fd. variable (Millen 20100).
C_4, C_5, C_7, C_9, C_{10} — 2–30-$\mu\mu$fd. ceramic trimmer (Millen 27030).
C_6 — 33-$\mu\mu$fd. silver mica.
C_8 — 10-$\mu\mu$fd. silver mica.
C_{11}, C_{12} — 680-$\mu\mu$fd. silver mica.
C_{13} — 68-$\mu\mu$fd. silver mica.
C_{14}, C_{15}, C_{17}, C_{18}, C_{19}, C_{21}, C_{22}, C_{23}, C_{24}, C_{27} — 0.01-μfd. 400-volt paper.
C_{16}, C_{20} — 100-$\mu\mu$fd. mica.
C_{25}, C_{26} — 47-$\mu\mu$fd. mica.
R_1, R_9 — 0.1 megohm, ½ watt.
R_2, R_{10} — 10,000 ohms, ½ watt.
R_3 — 47 ohms, ½ watt.
R_4 — 330 ohms, 1 watt.
R_5 — 15,000 ohms, 2 watts.
R_6 — 1 megohm, ½ watt.
R_7, R_{13} — 0.22 megohm, ½ watt.
R_8 — 0.5-megohm potentiometer.
R_{11} — 0.47 megohm, ½ watt.
R_{12} — 470 ohms, ½ watt.
R_{14} — 7500 ohms, 10 watts.
L_1 — 24 turns No. 22 tinned wire, diameter 1½ inches, length 1⅛ inches (B & W 80 JCL with 18 turns removed).
L_2, L_3 — 14 turns No. 24 e. wire, diameter 1 inch, length ⅜ inch; wound on Millen 45000 form.
L_4 — 3 turns No. 24 e., close-wound at bottom end of L_3.
J_1, J_2 — Coaxial-cable jack (Jones S-101).
RFC_1, RFC_3 — 2.5-mh. r.f. choke (Millen 34100).
RFC_2 — 300-μh. r.f. choke (Millen 34300).
S_1 — 4-position progressive-shorting switch (Centralab GG modified; see text).
S_2 — S.p.s.t. toggle switch.

swing is useful on 144 Mc., where a considerable number of relatively-broad receivers is in use. The deviation is controllable to any required value below this, by means of the potentiometer, R_8. A switch is provided in the heater circuit of the speech section (S_2) so that this portion of the unit can be cut off when c.w. or amplitude modulation is being used. As operation of this switch affects the oscillator frequency appreciably it is usually preferable to leave the speech-section heaters on at all times, using the deviation control at its off position when emissions other than NFM are being used.

The arrangement of the parts should be clear from the photographs. The top view, Fig. 13-8, shows the microphone jack and heater switch at the right end of the panel. The deviation control, bandswitch, oscillator-plate and amplifier-plate tuning controls are in line across the bottom of the panel. The oscillator frequency setting is controlled by the vernier dial. Looking at the top of the chassis the two 6AG7s may be seen to the left of the tuning condenser, the first being the oscillator tube. The oscillator tank coil, L_1, is mounted on stand-offs, just in back of the 6AG7s. Two metal brackets are used to mount the tuning condenser, which should be the double-ended variety for greatest mechanical stability. The reactance-modulator and speech-amplifier tubes are at the right of the tuning condenser, with the regulator at the rear. The chassis is a standard 3 × 5 × 10-inch size and the panel is

6 by 11 inches. A 5 × 10-inch aluminum plate, with clearance holes for the trimmer adjustments, is attached to the bottom of the chassis.

The arrangement of components under the chassis is apparent from the bottom view, Fig. 13-10. The bandswitch and associated padders are at the middle, with the oscillator plate coil. The amplifier plate coil is at the left. The padder condensers are mounted with their grounded terminals soldered to metal pillars, in order to reduce sensitivity to vibration to a minimum.

The bandswitch requires some modification. In its original form it has a disk which shorts out all unused contacts. This disk must be cut through the center so that one half may be removed. As may be seen from the wiring diagram, Fig. 13-9, the connection between the oscillator coil and the switch is made to Number 1 terminal, rather than to the regular wiper contact.

The power supply for the VFO should be well filtered and capable of delivering 300 volts d.c. at 60 to 70 ma., and 6.3 volts a.c. at 1.9 amp. Socket voltage measurements are approximately as follows: 20 volts on the audio-tube screens, 150 volts on the 6AG7 screens, 40 and 150 volts, respectively, on the speech-amplifier and reactance-modulator plates, and 300 volts on the 6AG7 plates. Cathode current for the oscillator should be about 10 ma., and the output stage, at resonance, 30 ma.

Calibration and Use

Calibration of the VFO dial can be accomplished with the aid of a receiver having an accurate dial calibration, as the readings on the VFO dial should not be relied upon for band-edge operation. The 50-Mc. range, requiring the least padder capacitance, should be calibrated first. Padders C_9 and C_{10} set at nearly full capacitance will provide the correct tuning range, which should be approximately 55 divisions spread over the middle of the dial scale. The 144-, 28- and 27-Mc. ranges should be calibrated in that order, their spread on the dial being approximately 25, 80 and 20 divisions respectively. If the NFM portion of the unit is to be used extensively it is recommended that the calibration procedure be carried out with the reactance-modulator heater on, as this tube affects the calibration appreciably.

When adjusting the plate circuits of the oscillator and amplifier stages it is recommended that the approximate settings of these controls for the middle of the band in question be marked on the panel. It will then not normally be necessary to readjust these controls when shifting frequency within a band. This broad-band effect is accomplished by slightly overdriving the amplifier tube at the center frequency, causing the screen voltage to drop and reduce the output. Tuning away from the center frequency reduces the drive and allows the screen voltage and output to rise. More than enough output is thus obtainable over the entire band, without too great a variation for proper operation of the succeeding stage. Two 250-ma. pilot lamps in parallel make a satisfactory dummy load for the amplifier.

Next the operation of the reactance modulator should be checked. The procedure for this operation is described in detail in Chapter Nine. It should also be pointed out that there is no excuse for radiation of an improperly-modulated FM signal, since it can be monitored readily in one's own receiver. With the receiver in operation on the band in which the transmitter is to be used, but with only the VFO turned on, it is a simple matter to tell exactly how the signal will sound on the air. Deviation requirements vary with different receivers, but a safe starting point is to set the deviation control so that the signal sounds well on a communications receiver with the crystal filter in the broadest "on" position.

Ordinarily a unit of this type may be used to replace the crystal stage of an existing transmitter by simply plugging it into the crystal socket. The output coupling is a low-impedance line, however, and it may be connected to a link winding on the grid coil of any low-power stage whose tuning range is 7 to 9 Mc. Although it is shown calibrated only for the frequencies above 27 Mc., it may be used as a c.w. exciter for 7- or 14-Mc. work. The deviation may, however, be insufficient for 20-meter NFM operation. Output, at 7 to 9 Mc., is about three watts.

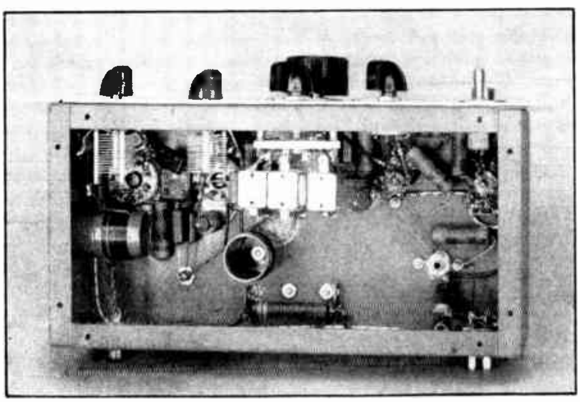


Fig. 13-10 — Bottom view of the VFO.

A Simplified Exciter for 50 and 144 Mc.

Through the use of a special crystal-oscillator circuit, by means of which standard low-cost crystals are made to oscillate on their third harmonic in a simple triode regenerative oscillator, the transmitter-exciter shown in Figs. 13-11, 13-12 and 13-13 provides output on 50 and 144 Mc. with only two tubes and simple circuits. A dual-triode oscillator-multiplier is used, the first section oscillating on 24 to 27 Mc., depending on the frequency of the crystal, which may be anything from 8 to 9 Mc. The second section doubles to 48 to 54 Mc., providing more than enough output to drive an 832 amplifier or tripler. Plug-in coils are used in the 832 plate circuit, to permit output on 50–54 Mc. or 144–148 Mc. Output on the lower band is 20 watts or more, with three to five watts available on the higher frequency.

The rig may be modulated on 50 Mc., in either portable or fixed-station service, but it should be used as an exciter only for 144 Mc. The power output on the higher band is sufficient to drive another 832 or 829 stage, the design of which might follow that of the 829 amplifier described later in this chapter. It may also be used to drive a 50-Mc. amplifier such as that in Fig. 13-6.

A standard 5 × 10 × 3-inch chassis is used, with the oscillator-multiplier components mounted below the deck and the 832 plate-circuit components above. A power-switching arrangement is included to permit use of the rig as a complete transmitter for 50 Mc. or as an exciter for an additional modulated stage on 144 Mc.

The Harmonic-Oscillator Circuit

Design is conventional except for the oscillator circuit, the key feature of which is the feed-back arrangement in L_1. The portion of the coil below the tap determines the proper functioning of the oscillator, the correct position of the tap being approximately one-third up from the crystal end of the coil when a 6J6 is used. With other dual triodes it may be necessary to alter this materially

If too much inductance is included in the tickler portion of the coil the tube will oscillate at a frequency determined by the setting of C_2 rather than by the crystal. When the unit is ready for test the oscillator stage alone should be checked first. With a low-range milliammeter inserted temporarily in series with the multiplier grid resistor, R_2, about 150 volts should be applied to the oscillator plate. Rotate C_2 until grid current appears, indicating oscillation, the frequency of which should be checked in a calibrated receiver. Changing the setting of C_2 should not cause an appreciable change in the frequency of oscillation, and the crystal will oscillate only over a part of the tuning range of the condenser and at no other point. If the oscillator frequency shifts widely, indicating uncontrolled oscillation, the tap is too high on L_1. If the tap is too low the 6J6 will oscillate weakly or not at all, and will refuse to start when the condenser is tuned near the point of maximum output, as indicated by the grid-current peak in the succeeding stage.

It should be noted that pulling the crystal out of its socket is *not* a satisfactory check for uncontrolled oscillation, as the capacitance of the crystal and its holder is required to complete the feed-back circuit.

Provision is made for measuring the grid and cathode current of the amplifier stage by means of J_1 and J_2. The former is insulated from the panel, and connected in reverse, so that the meter leads need not be reversed in changing from one jack to the other. When the rig is operated on 50 Mc. the grid current in the 832 need not be more than 2 ma., and this amount of drive can be furnished by the 6J6 with 150 volts applied to the junction of R_1 and R_4. Amplifier cathode current, with no load, will be about 35 ma. at resonance, with a 400-volt supply. It may be loaded up to about 70 ma.

If 144-Mc. output is desired, the final stage should not be operated at more than 300 volts or so, but at this level it will provide more than

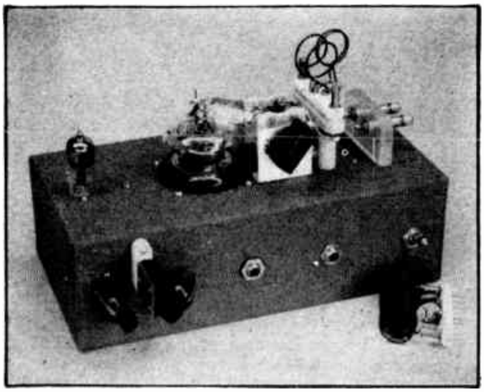


Fig. 13-11 — The two-tube exciter for 50 and 144 Mc. The 2-meter coil is plugged into the output stage, with the 6-meter one in the right foreground.

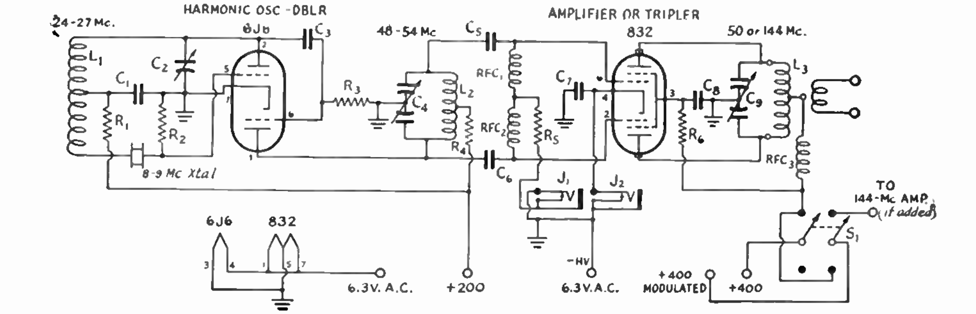


Fig. 13-12 — Schematic diagram of the 2-tube v.h.f. rig. The power-switching arrangement shown provides for later addition of a 144-Mc. amplifier stage.

C_1 — 680-$\mu\mu$fd. mica.
C_2 — 50-$\mu\mu$fd. variable.
C_3 — 15-$\mu\mu$fd. ceramic.
C_4 — 20-$\mu\mu$fd.-per-section split-stator, made by sawing the stator bars of a Millen 21050 and removing center plate.
C_5, C_6 — 75-$\mu\mu$fd. ceramic.
C_7, C_8 — 500-$\mu\mu$fd. ceramic.
C_9 — 6-$\mu\mu$fd.-per-section split-stator (Millen 21906D).
R_1 — 4700 ohms, ½ watt.
R_2 — 3300 ohms, 1 watt.
R_3 — 47,000 ohms, ½ watt.
R_4 — 3300 ohms, 1 watt.
R_5 — 22,000 ohms, 1 watt.
R_6 — 25,000 ohms, 10 watts.
L_1 — 14 turns No. 18, ½-inch diam., 1 inch long, tapped at 4½ turns.
L_2 — 12 turns No. 18, ½-inch diam., ⅞ inch long, center-tapped.
L_1 and L_2 made from Barker and Williamson "Miniductor" type 3003.
L_3 — 50 Mc. — 14 turns No. 14 enamel, ⅞-inch diam., 2 inches long. Link: 3 turns No. 20 enamel, spaghetti-covered.
144 Mc. — 2 turns No. 14 enamel, 1-inch diam., spaced ½ inch. Link: 2 turns No. 16 enamel.
Base and plug assemblies are National XB-16 and PB-16.
J_1, J_2 — Closed-circuit jack.
RFC_1, RFC_2, RFC_3 — 25 turns No. 24 enamel on 1-watt resistor, or Millen 34300.
S_1 — D.p.d.t. toggle switch.

enough output to drive another 832 amplifier, or even an 829. For 144-Mc. use the whole unit may be operated from a single 300-volt supply, the additional voltage on the oscillator and doubler being helpful in securing sufficient drive to make the 832 triple effectively. It is not recommended that the 832 be modulated for 144-Mc. voice operation, as there is not enough drive for operation of the stage as a modulated tripler, and the functioning of such a stage would not be generally satisfactory under any conditions. Grid current, for tripling, should be 4 ma. or more.

In 50-Mc. service the over-all drain, with a 300-volt supply, is only about 85 to 90 ma., and under these conditions the amplifier delivers an output of about 10 watts, with a total load on the supply of less than 30 watts. On 144 Mc. the output is three to five watts.

A more complete description of the transmitter and the regenerative oscillator circuit used may be found in *QST* for October and November, 1948. The same technique could be employed to advantage in the construction of an exciter unit for 220 Mc., except that the second section of the 6J6 would be operated as a tripler to 75 Mc., instead of as a doubler to 50 Mc. More than enough output would be available to drive another 6J6 as a tripler from 75 to 225 Mc.

Another possibility in connection with the oscillator circuit used in this transmitter involves taking off the fifth harmonic instead of the third. Many 7-Mc. crystals can be used in this way, taking off the 5th harmonic from the first triode section, and then doubling in the second. Only an additional doubler stage is then needed to reach 144 Mc.

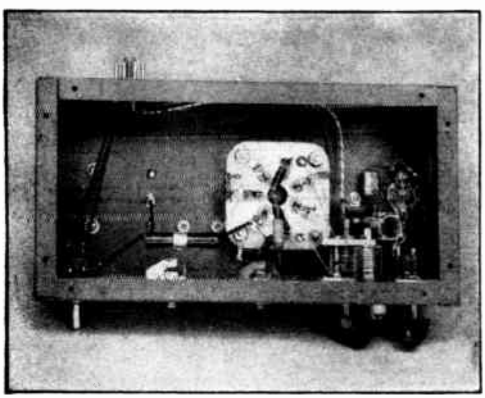


Fig. 13-13 — Bottom view of the simplified v.h.f. exciter.

144-Mc. Double Beam-Tetrode Power Amplifier

An amplifier set-up suitable for use with double beam-tetrode tubes is shown in Figs. 13-14, 13-15 and 13-16. The tube in the photographs is an 829, but an 815 or 832 can be used in the same layout. The only change that might be required would be in the inductances of the grid and plate coils, L_2 and L_3; these may have to be made slightly smaller or larger in diameter to compensate for the differences in input and output capacitances in the various types. When an 829 is used, the amplifier is well suited for use as an outboard unit with war-surplus transmitters such as the SCR-522.

The amplifier is built on an aluminum chassis formed by bending the long edges of a 5 × 10-inch piece of aluminum to form vertical lips ¾ inch high, so that the top-of-chassis dimensions are 3½ by 10 inches. The tube socket is mounted on a vertical aluminum partition measuring 3½ inches high by 3¼ inches wide on the flat face, with the sides bent as shown in the photographs to provide bracing. The partition is mounted to the chassis by right-angle brackets fastened to the sides. The socket is mounted with the cathode connection at the top, the cathode prong being directly grounded to the nearest mounting screw for the socket. The heater by-pass condenser, C_6, is mounted directly over the center of the tube socket, extending between the paralleled heater prongs at the bottom and the cathode prong at the top. The screen by-pass is connected with short leads between the screen prong and the nearest socket screw.

The grid coil, L_2, is supported by the grid prongs on the socket. The two turns of the coil are spaced about one-half inch to allow room for the input coupling coil L_1 to be inserted between them. The coupling is adjusted by bending L_1 into or out of L_2. The grid tuning condenser, C_1, is mounted between the socket prongs; although the condenser has mica insulation it is used essentially as an air-dielectric condenser since the movable plate does not actually contact the mica at any setting inside the band. The coupling link is soldered to lugs under binding posts on a National FWG strip, the strip being mounted on metal pillars 1½ inches high to bring the link to the same height as the grid coil.

Although the shielding between the input and output circuits of the tube is sufficiently good so that the circuit will not self-oscillate, tuning of the plate circuit will react on the grid circuit to some extent because the grid-plate capacitance, while small, is not zero. To eliminate this reaction it is necessary to neutralize the tube. The neutralizing "condensers" are lengths of No. 12 wire soldered to the grid prongs on the socket. The wires are crossed over the socket and then go through small ceramic feed-throughs at the top of the vertical shield, projecting over the tube plates on the other side as shown in Fig. 13-14.

Connections between the plate tank condenser, C_7, and the tube plate terminals are made by means of small Fahnestock clips soldered to short lengths of flexible wire. The tank coil, L_3, is mounted on the same condenser terminals to which the plate clips make connection. The output link, L_4, is mounted similarly to the grid link except that the posts are 1⅞ inches high. The plate choke, RFC_1, is mounted vertically on the chassis midway between the plate prongs of the tube, the mounting means being a short machine screw threaded into the end of the polystyrene rod. The "cold" lead of the choke is by-passed by C_5 underneath the chassis.

Supply connections are made through a 5-post strip on the rear edge of the chassis. The dotted lines between connections in Fig. 13-15 indicate

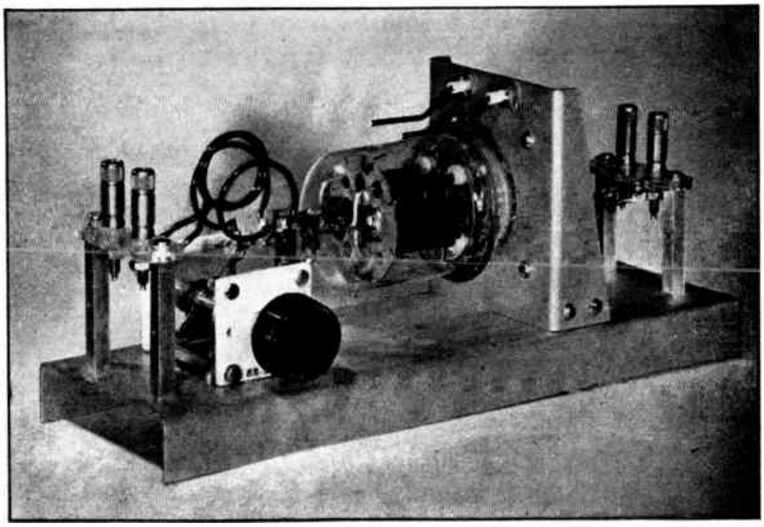


Fig. 13-14 — A 144-Mc. amplifier using a double beam tetrode. This type of construction is suitable for the 815 and 832 as well as the 829 shown. The vertical partition provides support for the tube as well as shielding between the input and output circuits. Note the neutralizing "condensers" formed by the wires near the tube plates.

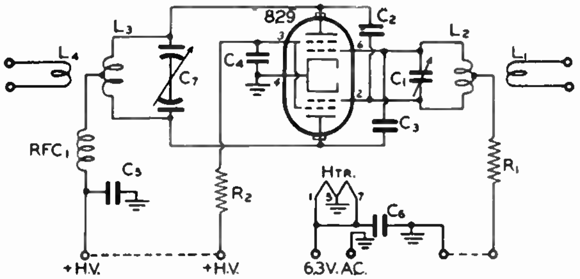


Fig. 13-15 — Circuit of the 829 amplifier for 144 Mc.

C_1 — 3–30-μμfd. ceramic trimmer.
C_2, C_3 — Neutralizing condensers; see text.
C_4 — 500-μμfd. mica, 1000 volts.
C_5 — 500-μμfd. mica, 2500 volts.
C_6 — 470-μμfd. mica.
C_7 — Split stator, 15 μμfd. per section (Cardwell ER-15-AD).
R_1 — 4700 ohms, 1 watt.
R_2 — 10,000 ohms, 10 watts.
L_1 — 2 turns No. 12, diameter ½ inch.
L_2 — 2 turns No. 12, diameter ½ inch, length ½ inch.
L_3 — 2 turns No. 12, diameter 1⅛ inches, length 1 inch.
L_4 — 2 turns No. 12, diameter 1 inch.
RFC_1 — 1-inch winding of No. 24 d.s.c. or s.c.c. on ¼-inch diameter polystyrene rod.

that these connections are normally short-circuited; leads are brought out so that the grid and screen currents can be measured separately.

In adjusting the amplifier, the plate and screen voltages should be left off and the d.c. grid circuit closed through a milliammeter of 0–25 or 0–50 range. The driver should be coupled to the amplifier input circuit through a link (Amphenol Twin-Lead is suitable, because of its constant impedance and low r.f. losses). Use loose coupling between L_1 and L_2 at first, and adjust C_1 to make the grid circuit resonate at the driver frequency, as indicated by maximum grid current. The coupling between L_1 and L_2 may then be increased to make the grid current slightly higher than the rated load value for the tube used — approximately 12 ma. for the 829. If the driver is an oscillator, the coupling between L_1-L_2 should be as loose as possible with proper grid current.

After neutralization, the procedure for which has been given in connection with other similar amplifiers, plate and screen voltage may be applied. If possible, the plate voltage should be low at first trial so there will be no danger of overloading the tube. Adjust C_7 to resonance, as indicated by minimum plate current (this should be measured independently of the screen); with the 829, the minimum plate current should be in the neighborhood of 80 milliamperes with 400 volts on the plate and no load on the circuit. A dummy load such as a 60-watt lamp should light to something near full brilliance when the coupling between L_3 and L_4 is adjusted to make the tube draw a plate current of 200 ma. When the loading is set, the grid current should be checked to make sure it is up to the rating for the tube.

Power-supply and modulator requirements will depend upon the particular tube used. For the 829, the plate supply should have an output voltage of 400 to 500 with a current capacity of 250 milliamperes. With a 400-volt supply the modulator power required is 50 watts, with an output transformer designed to work into a 1600-ohm load; with a 500-volt supply slightly over 60 watts of audio power is needed, the load being 2000 ohms.

This amplifier may also be used with the 832 rig described in the preceding pages. The output from the driver stage may be fed into the amplifier by means of a link, if the two units are to be operated remote from one another, or the grid circuit of the 829 may be arranged to provide direct inductive coupling to the 832, if the two are placed side by side.

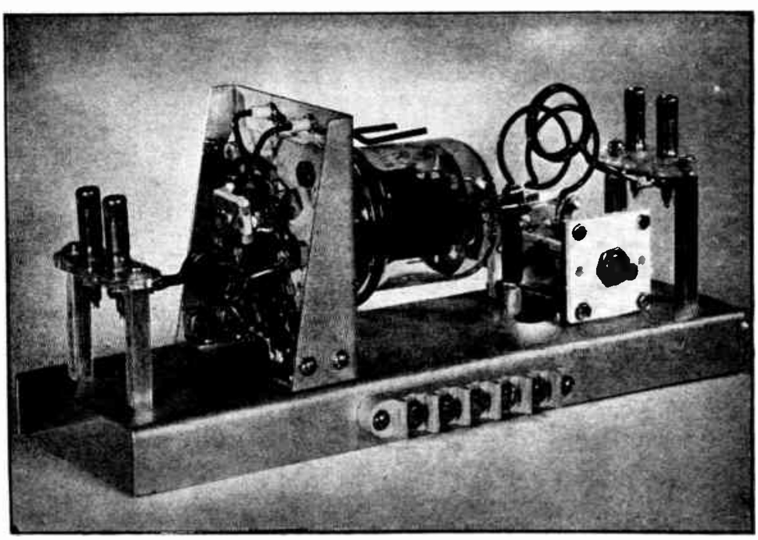


Fig. 13-16 — Another view of the 144-Mc. amplifier. The neutralizing wires are crossed over the socket before going through the feed-through insulators. The input circuit is designed for link coupling to the driver stage.

Crystal Control on 220 Mc.

Construction of a multistage transmitter for the 220-Mc. band is not as difficult as might be imagined, and the serious worker on this frequency will find the use of crystal control or its equivalent highly worth while. Fortunately the crystals used are also usable on 144 Mc., cutting down the total cost of building equipment for both bands, if the crystal frequencies are selected with this use in mind.

The transmitter-exciter shown in Figs. 13-17, 13-18 and 13-19 employs either 8- or 12-Mc. crystals, and if they are between 8148 and 8222 or 12,223 and 12,333 kc. they may also be used for operation in the upper portion of the 144-Mc. band. By using miniature tubes and components, and by arranging the parts for minimum lead length, efficient operation on 220 Mc. is obtained, with a simplicity of construction that puts the equipment well within the capabilities of the average experienced amateur.

Four 6J6 dual triodes are used. The first works as a triode oscillator and frequency multiplier, the second section doubling or tripling, depending upon which type of crystal is employed. Tuning is less critical, and the various stages operate somewhat more efficiently with 12-Mc. crystals, but 8-Mc. crystals may also be used. The next two stages are push-pull triplers, and the output stage is a neutralized amplifier. Capacitive coupling is used between stages. The chassis is 2½ inches wide, 2 inches high, and 12 inches long, with ½-inch edges folded over. It may be made from a piece of sheet aluminum 7½ by 12 inches in size. The first tube socket is 1½ inches in from the left end and the other sockets are spaced along the chassis, 2¼ inches center to center. The tuning condensers are spaced equally between the sockets, the last two, C_{13} and C_{17}, being mounted on the top surface of the chassis for minimum lead length and symmetrical layout. Pin jacks, labeled *a* and *b* on the schematic diagram, are mounted on the front wall of the chassis and may be used for metering or keying of the output stage.

Initial Adjustments

Meter jacks for the individual stages were not considered necessary, as there will normally be few occasions for shifting frequency and retuning, once the initial adjustment of the exciter is completed. For these first measurements the various circuits may be opened and tests made with a portable meter.

With a meter in series with R_2, set the core in L_1 at an intermediate position and adjust C_2 for oscillation, as indicated by a dip in plate current to about 10 ma. The frequency and note should be checked in a communications receiver, making sure that the oscillation is controlled by the crystal. Next, insert the meter in series with R_4 and tune C_4 for a dip at the proper frequency, which should be between 24.5 and 25 Mc. Adjustment of the multiplier tuning may be critical, if fundamental-type crystals are used, the crystal tending to "pop out" when C_4 is tuned on the nose. With "overtone" or harmonic-type crystals this trouble will not be in evidence, and the setting of C_4 (or the core in L_2) will not be fussy. Adjustment should be for maximum grid current in the second 6J6.

Adjustment of the push-pull tripler stages is merely a matter of resonating the circuits for maximum output as indicated by the grid current in the succeeding stage, being certain that the stages are tripling and not quintupling, which they will also do with fair efficiency. Each stage has cathode bias to prevent damaging the tubes during the adjustment period. Input to each will run about 25 ma. at 200 volts, when operating correctly.

Neutralization of the output stage is accomplished in the customary manner, except that the neutralizing capacitors are made from short lengths of 75-ohm Twin-Lead.

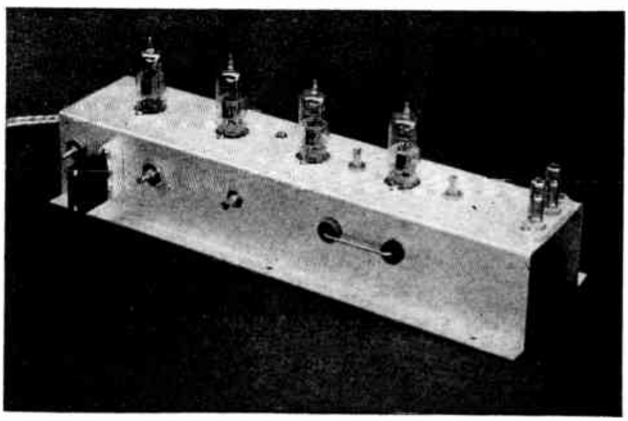


Fig. 13-17 — Front view of the 220-Mc. transmitter-exciter. Across the front of the chassis are the oscillator plate-coil adjustment, crystal, multiplier-coil adjustment, first-tripler plate condenser, and tip jacks for final cathode metering. Second-tripler and final plate condensers are mounted on the top portion of the chassis. Output terminals are at the far right.

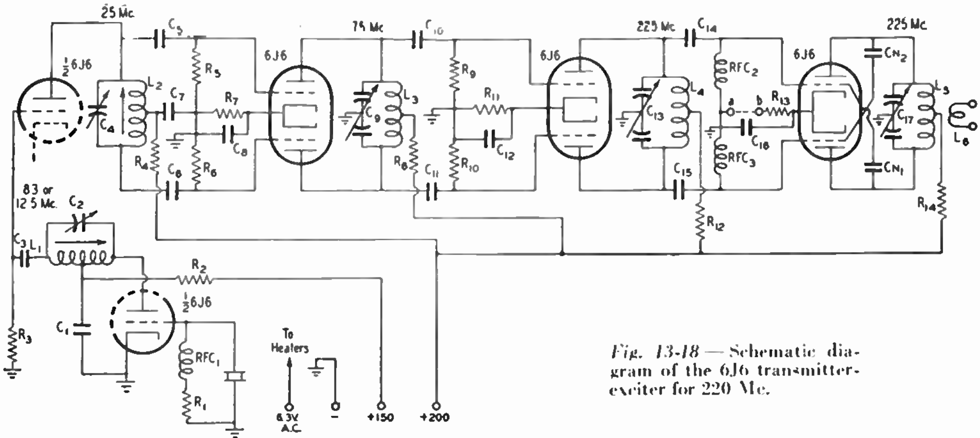


Fig. 13-18 — Schematic diagram of the 6J6 transmitter-exciter for 220 Mc.

C_1, C_7 — 680-μμfd. mica.
C_2, C_4 — 3–30-μμfd. mica trimmer.
C_3 — 68-μμfd. mica.
C_5, C_6 — 47-μμfd. mica.
C_8, C_{12} — 330-μμfd. mica.
C_9, C_{13} — 2.7–8.5-μμfd. midget butterfly variable (Johnson 160–208).
C_{10}, C_{11}, C_{14}, C_{15} — 50-μμfd. ceramic (National XLA-C).
C_{16} — 200-μμfd. ceramic.
C_{17} — 1.7–3.3-μμfd. midget butterfly variable (Johnson 160–203).
C_{N1}, C_{N2} — Neutralizing capacitors made of 75-ohm Twin-Lead: see text.
R_1, R_3 — 6800 ohms, ½ watt.
R_2 — 470 ohms, ½ watt.
R_4 — 3900 ohms, 1 watt.
R_5, R_6, R_9, R_{10} — 22,000 ohms, ½ watt.
R_7, R_{11}, R_{13} — 470 ohms, 1 watt.
R_8, R_{12}, R_{14} — 1500 ohms, 1 watt.
L_1 — 34 turns No. 28 d.s.c., close-wound on National XR-50 slug-tuned form, center-tapped.
L_2 — 12 turns No. 21 d.s.c., close-wound on National XR-50 slug-tuned form, center-tapped.
L_3 — 7 turns No. 16 enamel, ⅝-inch inside diameter, spaced wire diameter, center-tapped.
L_4 — 2 turns No. 16 enamel, ⅜-inch inside diameter, spaced ¼ inch, center-tapped.
L_5 — 1½ turns No. 12 enamel, ¾-inch inside diameter, center-tapped. Space turns about 3/16 inch apart. Coil 1½ inches long over-all. See bottom-view photograph.
L_6 — Hairpin loop No. 16 enamel inserted between turns of L_5.
RFC_1 — 250-μhy. r.f. choke (Millen 34300).
RFC_2, RFC_3 — Solenoid v.h.f. choke — No. 28 d.s.c. wire wound on ½-watt carbon resistor, ⅛-inch diameter, 9/16 inch long.

Starting with sections about two inches long, they should be trimmed a small amount at a time until tuning the final plate through resonance (with plate voltage removed) causes no downward kick in grid current.

Performance

With the voltages shown, the output on 220 Mc. will be about 2 watts, as indicated by a full-brilliance indication in a Number 46 (blue bead) pilot lamp. More output can be obtained by increasing the voltage above 200, but the increase is seldom worth the extra strain on the tubes. Operated as shown, the rig will give ample output to drive an 832 amplifier which will deliver about 12 watts, or the final 6J6 may be modulated and the unit operated as a complete low-powered transmitter.

The same general arrangement described above may be used to get to 220 Mc. with three tubes instead of four, if the regenerative harmonic-oscillator circuit shown in Fig. 13-12 is used to replace the more conventional crystal oscillator circuit of Fig. 13-18. An 8.3-Mc. crystal is then made to oscillate on 25 Mc. in the first 6J6 section. The second section triples to 75 Mc. The rest of the unit, from L_3 on, is the same as in Fig. 13-18. It is suggested that the description of the 6- and 2-meter transmitter of Fig. 13-12 be studied carefully before this substitution is attempted.

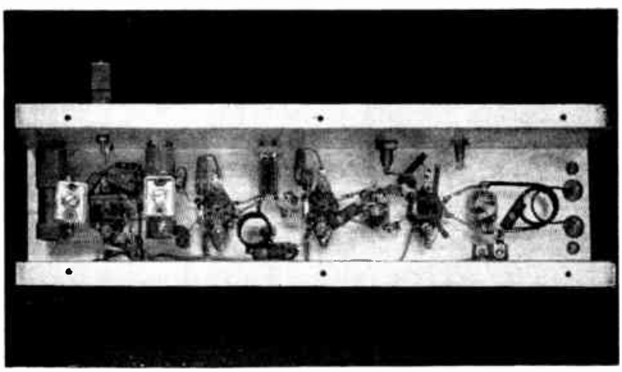


Fig. 13-19 — Bottom view of the 6J6 220-Mc. rig, showing the simplicity of the layout.

Simple Gear for 144 and 220 Mc.

Until recently, most stations operating in the higher v.h.f. bands employed simple transmitters of the modulated-oscillator type. Since the superregenerative receiver was also widely used, the instability of the transmitters was not a matter of great importance; but with the rapid swing to stabilized transmitters and selective receivers now in evidence, most of the modulated-oscillator signals are no longer readable. It is, however, still possible, by careful design and proper operation, to use the simple and economical oscillator rig and yet radiate a signal that can be copied on all but the most selective receivers. Two such transmitters, for 144 and 220 Mc., are shown in Figs. 13-20 through 13-27.

Oscillator Ills and Their Treatment

There are two principal faults in most simple oscillator-type transmitters. Many use filament tubes with a.c. applied to the filaments, causing severe hum modulation. Others, through poor design, have insufficient feedback (as evidenced by low grid current) so that they are unable to sustain strong oscillation under load. Lack of sufficient excitation also renders them incapable of maintaining oscillation at low plate voltages, causing them to go out of oscillation over a considerable portion of the modulation cycle. Such oscillators suffer from extreme frequency modulation, making their signals unreadable on all but the very broadest receivers, and even on these the quality is poor indeed.

A 2-METER UNITY-COUPLED OSCILLATOR

No simple transmitter can hope to overcome these faults entirely, but they are materially reduced in the rig described herewith. A.c. hum modulation is reduced through the use of indirectly-heated tubes; and stability is improved through the use of a high-*C* push-pull oscillator, employing the familiar "unity-coupled" circuit. This arrangement, wherein the grid coil is fed through the inside of a plate tank made of copper tubing, provides adequate excitation. Stability over wide ranges of plate voltage is quite good, and the degree of frequency modulation is not too severe if the modulation is held to 75 per cent or less. It is laid out so that it is stable mechanically, reducing possible frequency changes from vibration.

Mechanical Details

The transmitter is designed for use with a plate supply of 250 to 300 volts, making it useful for mobile or low-powered home-station

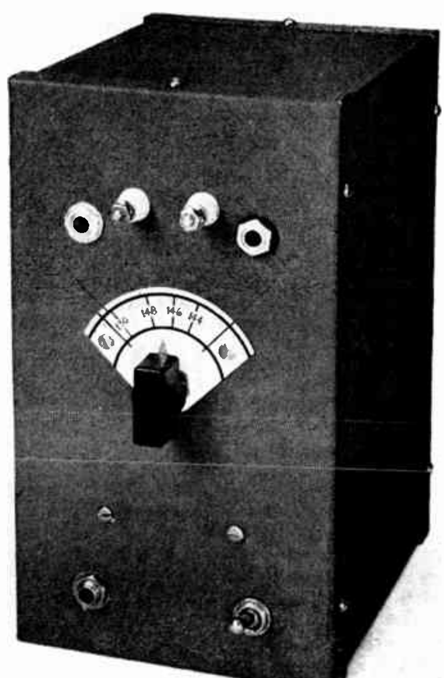


Fig. 13-20 — Front view of the simple 144-Mc. transmitter. The jacks at each side of the antenna terminals are for insertion of a meter in the oscillator grid (left) and plate (right) circuits. The microphone jack is at the lower left and the on-off switch is at the right. The calibration scale is drawn with India ink on heavy white paper.

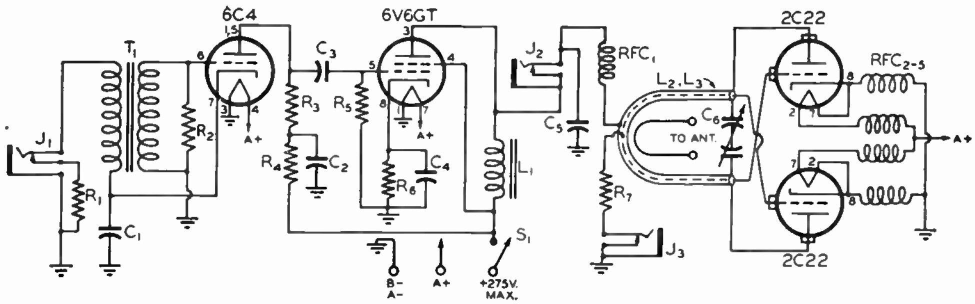


Fig. 13-21 — Schematic diagram of the simple 144-Mc. transmitter.

C_1, C_4 — 10-μfd. 25-volt electrolytic.
C_2 — 8-μfd. 450-volt electrolytic.
C_3, C_5 — 0.01-μfd. 600-volt paper.
C_6 — "Butterfly" variable (Cardwell ER-14-BF/S modified; see text).
R_1 — 470 ohms, 1 watt.
R_2 — 0.33 megohm, ½ watt.
R_3, R_4 — 5000 ohms, 5 watts.
R_5 — 0.47 megohm, ½ watt.
R_6 — 680 ohms, 1 watt.
R_7 — 10,000 ohms, 1 watt.
L_1 — Midget filter choke.
L_2, L_3 — Unity-coupled grid and plate coils. See text and Fig. 13-22.
J_1, J_2, J_3 — Closed-circuit jack.
RFC_1 — No. 28 d.s.c. wire, close-wound on 1-watt resistor, ¼-inch diam., ⅝ inch long.
RFC_2, RFC_3, RFC_4, RFC_5 — 20 turns No. 20 d.s.c. wire close-wound on ¼-inch polystyrene rod.
S_1 — S.p.s.t. toggle switch.
T_1 — Single-button microphone transformer (UTC "Ouncer" — surplus).

operation. It employs a pair of 2C22 tubes (also known as 7193s) as oscillators, a 6V6GT modulator, and a 6C4 as a speech amplifier and source of microphone voltage. It is housed in a standard 5 × 6 × 8-inch utility cabinet, the back and front of which are removable. The schematic diagram is shown in Fig. 13-21.

The plate tank "coil" is made of 3/16-inch copper tubing, bent into a "U" which is two inches long overall. The ends of the "U" are made into spade lugs, as shown in Fig. 13-22, the slotted ends providing a small range of inductance adjustment. The lug ends are fastened directly to two of the stator terminals of the butterfly-type tank condenser, C_6. Part of the "U" is cut out at the curved end, to provide an opening for the center-tap of the grid coil. An easy way to make the grid coil is to cut two pieces of flexible insulated wire, about four inches long, and feed them into the "U" through the center opening. The protruding tap, made by twisting the ends of the wires together, should be coated with household cement after the grid resistor has been soldered to it. Note that the grid leads are transposed. The 2C22s will not oscillate if these are improperly connected. The plate leads may be made of ¼-inch copper braid, or copper or silver ribbon is even better, if available. If braid is used, it may be made solid at the end by flowing solder over the last half inch, after which it may be drilled, to pass the stator terminal screw.

Provision is made for reading both grid and plate current to the oscillator, two meter jacks being mounted on either side of the plate tank. Their terminals make convenient mounting places for R_7 and RFC_1. Note that the jacks are connected so that the meter leads need not be reversed when changing from one jack to the other. The plate-meter jack must, of course, be insulated from the metal panel.

No battery is required for microphone current, this being obtained by running the cathode current of the 6C4 speech amplifier through the microphone transformer. The 6C4 cathode is by-passed with a large electrolytic condenser, and the plate is decoupled and by-passed to reduce hum. Since the 6C4 stage is used principally as a source of microphone current, resistance coupling to the 6V6GT modulator gives adequate drive. No gain control is included, as the full output of the modulator is insufficient for overmodulation.

Testing

Since the grid is the controlling element in the operation of any Class C stage, it is important that the grid current be observed in adjust-

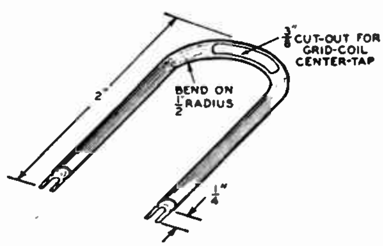


Fig. 13-22 — Detail drawing of the oscillator plate inductance. It is made from 3/16-inch copper tubing, bent into a "U" shape. Ends of the "U" are formed into spade lugs, the slots in which provide a means of slight inductance adjustment. It is mounted directly on the stator terminals of the tuning condenser.

ing the oscillator. The plate current may be almost meaningless, as an indication of the proper functioning of such a stage, but the grid current shows plainly if the oscillator is functioning correctly. If the grid current and bias are normal for the tubes used, the plate current can be ignored, except to see that the input is not excessive. Grid current in this os-

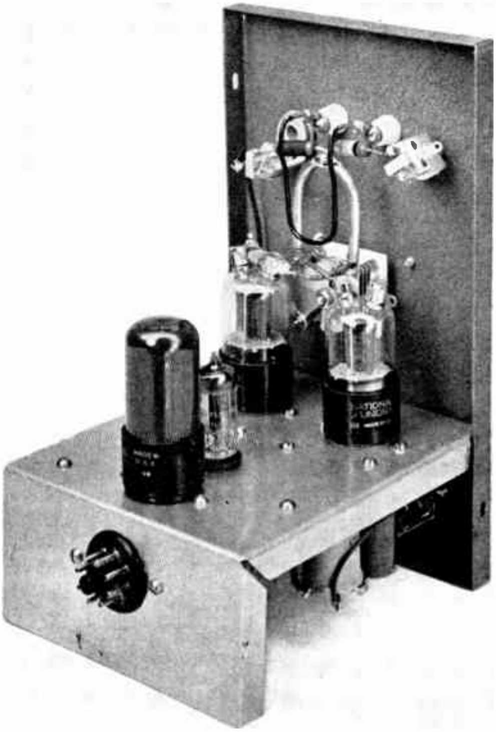


Fig 13-23 — Back view of the 2-meter transmitter, showing the symmetrical arrangement of components. Note that the "U"-shaped tank inductance is mounted directly on the stator terminals of the butterfly tuning condenser.

cillator should run about 8 ma. when a plate voltage of 275 or so is used and the oscillator is loaded by a lamp or antenna. The "U"-shaped antenna-coupling loop should be adjusted until the grid current is approximately this value. The plate current will be about 60 ma. with 275 volts on the plates.

The transmitter frequency should be checked with Lecher wires, or by listening to the signal in a calibrated receiver. In either case there should be a load across the antenna terminals, as the frequency may be appreciably different between loaded and unloaded operation.

The rough calibration scale shown was first roughed on a white card using pencil, and afterward drawn over in India ink. The calibration card is glued to the panel, and further held in place by the condenser mounting nut and two small machine screws.

A LINE OSCILLATOR FOR 220 MC.

A line oscillator which is suitable for low-power experimental work is shown in Figs. 13-25, 13-26 and 13-27. It is built entirely of readily-obtainable standard parts, and may be constructed at very low cost. The tube is a 7F8 dual triode, working as a push-pull oscillator, with parallel lines in the plate circuit. The frequency is varied by means of a mica trimmer which is connected across the line near the cold end, so that a vernier effect is attained. A rough adjustment of frequency is made by means of an adjustable shorting bar.

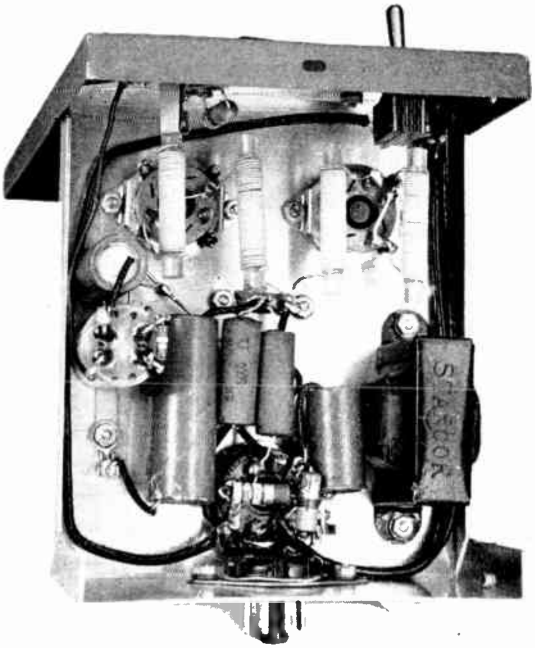


Fig. 13-24 — Under-chassis view shows the four heater chokes and audio components. The small round object, left center, is the microphone transformer, a surplus midget unit. The audio choke is at the right.

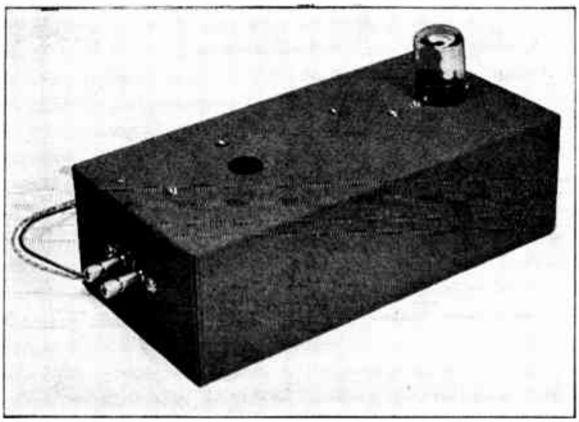


Fig. 13-25 — **A one-tube oscillator for 220 Mc. using a 7F8 dual triode. Linear tank circuit and antenna coupling are under the chassis.**

When the proper setting of the shorting bar is found, the 220–225-Mc. band will be covered by about two complete turns of the trimmer.

The transmitter is mounted on a 3 × 5 × 10-inch chassis. Only the oscillator tube is above the chassis, with the lines and antenna coupling below. The antenna coupling loop is connected to a National FWG terminal assembly which projects through the end of the chassis. The plate lines are 7½ inches long and made of ¼-inch copper tubing spaced ¾ inch, center to center. They are held in position by two halves of a National FWH or FWJ terminal block. These blocks are of low-loss insulating material, and the hole spacing is right for this application. The connection between the plates and the lines should be made with ¼-inch-wide copper strip. They are mounted on two cone stand-offs 6 inches apart. Self-supporting r.f. chokes, one in the cathode lead and the other in the B-plus lead, a 1000-ohm resistor from grids to ground, and a small by-pass condenser from the hot heater terminal to ground, complete the circuit. The antenna coupling is a "U"-shaped loop 4½ inches long.

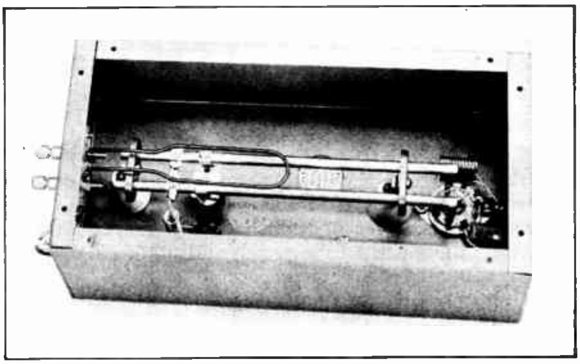


Fig. 13-26 — **Under-chassis view of the 220-Mc. transmitter. Note the method of making the shorting bar and mounting the trimmer condenser — both by the use of spring grid clips, permitting adjustment of the position of either along the line.**

The transmitter may be placed in operation by applying 6.3 volts a.c. and about 250 volts d.c. Plate current, under load, should be under 40 ma. A lamp load should be used across the antenna terminals until the frequency is adjusted to within the band limits. The shorting bar is made from two National No. 8 grid clips, which make a tight fit on the ¼-inch tubing, and the trimmer condenser is also connected to the line by means of a pair of these clips, making it possible to adjust the position of the condenser along the line to give the desired degree of frequency coverage. The shorting bar and the trimmer should be set in such positions that, with the trimmer set near maximum, the frequency of oscillation is near 220 Mc.

The antenna coupling may then be adjusted for maximum power transfer (a field-strength meter close to the antenna is a good indication), using the minimum coupling that will give satisfactory output. The frequency should be checked carefully with the antenna on and the coupling adjusted.

The transmitter can be run at 10 watts input without endangering the tube. The useful output is in the vicinity of 2 watts. The rig may be modulated with a single 6V6 tube, a suitable modulator being that shown in Fig. 13-21.

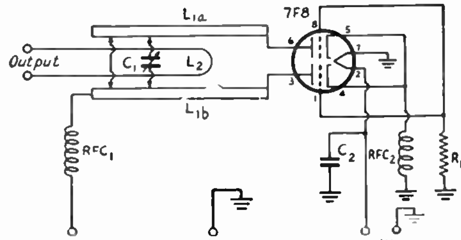


Fig. 13-27 — **Schematic of the 220-Mc. transmitter. R.f. chokes are 12 turns No. 18 d.c.c. wire, ¼-inch diam. C_2 is 0.0047 μfd. See text for other values.**

Mobile Gear with Quick-Heating Filaments

A worth-while saving in battery drain can be made by using filament-type tubes in the mobile station, arranging the control circuits so that the filament voltage is applied simultaneously with the starting of the generator or vibrator supply. The mobile transmitters shown in Figs. 13-28 to 13-36 combine operation on 50 and 144 Mc. They use Hytron instant-heating filament tubes throughout. All the necessary control and power-supply circuits are given in the schematic diagrams.

Fig. 13-28 shows the three units. At the left is the 144-Mc. transmitter, with the 50-Mc. rig at the right. The modulator, shown between them, may be used with either unit. By means of suitable interconnecting cables, connections for which are shown in the schematic diagrams, it is possible to select either band by operation of a single switch at the control position. Operation thereafter is controlled entirely by the push-to-talk switch on the microphone.

Both units use Valpey type CM-5 crystals in the 24–27-Mc. range, with a 2E30 Tri-tet oscillator doubling to 48–54 Mc. The oscillator-doubler drives a Hytron 5516 amplifier directly in the 50-Mc. transmitter. A Type 5812 tripler drives the 5516 final in the 144-Mc. rig. The modulator uses two 2E30s driven directly by a carbon microphone. Coaxial output fittings are provided for antenna connection, and a series-tuned antenna coupling circuit is included in each unit. Note that the jacks for metering purposes are recessed in back of the panels, to prevent contact with the high voltage, a danger spot in many mobile installations.

The 50-Mc. R.F. Section

The 50-Mc. r.f. unit, Figs. 13-29, 13-30, and 13-31, is built on an aluminum chassis 4 inches square and 2 inches high. The panel is 4 inches square, with a half-inch lip folded over across the bottom for fastening to the chassis. Arrangement of the parts is obvious from the photographs. It will be seen that the screen dropping resistor, R_2, is a lower value in this unit than in the 144-Mc. one. More oscillator power was required, as the final stage is driven directly, and the value of the screen resistor is a good means of controlling oscillator output.

No neutralization of the final was required, but a slight regenerative tendency at some condenser settings was corrected by the insertion of R_5, a 22-ohm resistor, at the grid terminal of the 5516.

The 144-Mc. Portion

The 2-meter r.f. section is built on a standard 2 × 5 × 7-inch chassis, with a 6 × 7-inch panel. The oscillator is similar to the 6-meter one, except as noted above. It is followed by a tripler stage using a 5812, a tube similar to the 2E30 but designed specifically for frequency multiplication. The plate circuit of this tube is inductively coupled to the final grid circuit, L_3 and L_4 being hairpin-shaped loops visible in the bottom view, Fig. 13-34.

Note the method of neutralization used in the final stage. The copper fin (designated as C_{16} in Fig. 13-33) visible in the rear view of the 144-Mc. unit is a device occasionally found necessary in tetrode amplifiers. In this

TABLE 13-II

Typical Operating Conditions in the 50- and 144-Mc. Mobile Transmitters of Fig. 13-28 When Used with a 300-Volt Supply.

Stage	*Plate Current*	*Screen Voltage*	*Grid Current*
50-Mc. Osc.	30 ma.	200 v.	—
144-Mc. Osc.	30	175	—
144-Mc. Tripler	40	150	—
50-Mc. Amp.	60	220	3 ma.
144-Mc. Amp.	60	160	3
Modulator	50–80	300	—

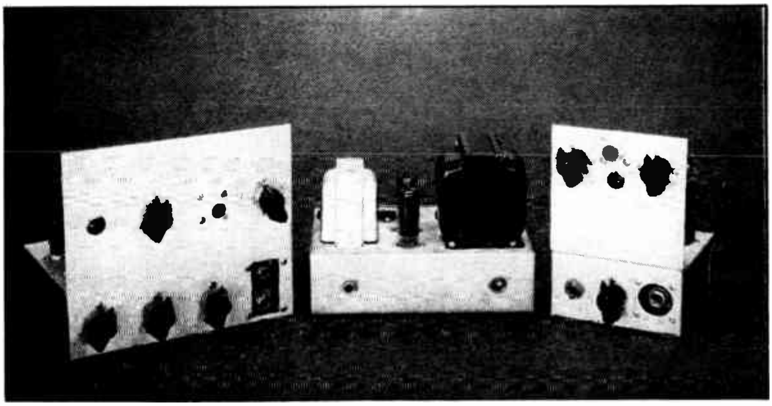


Fig. 13-28 — A complete mobile station for 50 and 144 Mc. using quick-heating filament tubes. The 144-Mc. r.f. section is at the left, the 50-Mc. portion at the right, and the modulator in the middle.

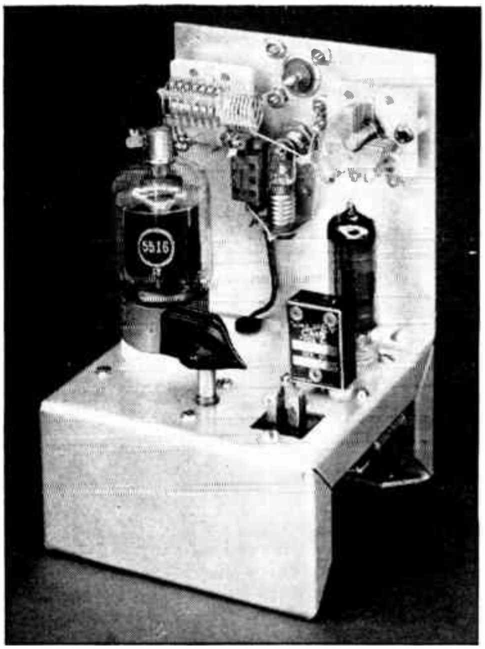


Fig. 13-29 — **Rear view of the 50-Mc. r.f. section. The knob above the chassis is the cathode control. The final tank circuit is at the upper left, with antenna series tuning at the upper right.**

case the physical layout was such that the grid-plate capacitance was effectively negative; thus the addition of external capacitance directly from grid to plate. The position of the fin is adjusted in the normal manner. It was made by hammering out the end of a piece of 3/16-inch copper tubing.

Details Common to Both Units

The Tri-tet circuit is modified for filament-type tubes by using closely-coupled (interwound) coils in the filament leads and tuning one of them. This cathode circuit is resonated slightly higher than the frequency marked on the crystal. It may be tuned for maximum grid current indication in the succeeding stage. There are various types of crystals for the 24–27-Mc. range. Until recently such crystals have been highly active but very unstable, and great care has been necessary to prevent extreme drift when they were used. Most crystal companies now supply harmonic-type crystals that are less active, but much more stable. The same cathode circuit will work with either variety, but more input will have to be run to the oscillator to achieve the same grid drive when the new type of crystal is used. If the old-type crystals are used the screen resistor, R_2, can be increased to as much as 120,000 ohms, dropping the total cathode current to about 20 ma. At this input the drift, with the unstable type of crystal, is not severe. It amounts to approximately 20 to 30 kc., at 144 Mc., but may be as much as ten times this value if the oscillator is not operated correctly. The newer types of crystals show a quick drift of a few kilocycles at 144 Mc., as the plate voltage is applied, but remain fairly steady after the first few seconds.

The cathode-circuit values given are correct for either type of crystal. The cathode coils, L_{1A} and L_{1B}, are made by winding with two wires simultaneously. A coating of household cement over the windings will hold them together, giving the coil the appearance of a single winding.

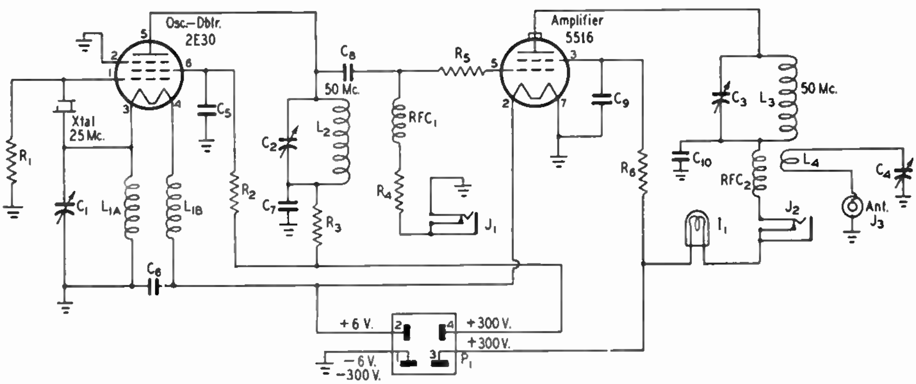


Fig. 13-30 — **Schematic diagram of the 50-Mc. mobile unit.**

C_1, C_4 — 50-μμfd. variable (Millen 20050).
C_2, C_3 — 15-μμfd. variable (Millen 20015).
C_5, C_6, C_7, C_9, C_{10} — 470-μμfd. mica.
C_8 — 22-μμfd. mica or ceramic.
R_1 — 0.1 megohm, ½ watt.
R_2 — 39,000 ohms, 1 watt.
R_3 — 100 ohms, ½ watt.
R_4 — 15,000 ohms, ½ watt.
R_5 — 22 ohms, ½ watt.
R_6 — 8000 ohms, 2 watts.
L_{1A}, L_{1B} — Interwound coils, each 12 turns No. 18 enamel, 3/8-inch diameter.
L_2 — 7 turns No. 18 tinned, ½-inch diameter, 7/8 inch long (B & W Miniductor, No. 3002).
L_3 — 8 turns No. 20 tinned, ½-inch diameter, 1 inch long (B & W No. 3002).
L_4 — 7 turns No. 20 tinned, ½-inch diameter, 7/16 inch long (B & W No. 3003).
I_1 — Pilot-lamp assembly with 60-ma. bulb.
J_1, J_2 — Closed-circuit jack.
J_3 — Coaxial output fitting.
P_1 — 4-prong male plug (Jones P-304-AB).
RFC_1, RFC_2 — 7-μh. r.f. choke (Ohmite Z-50).

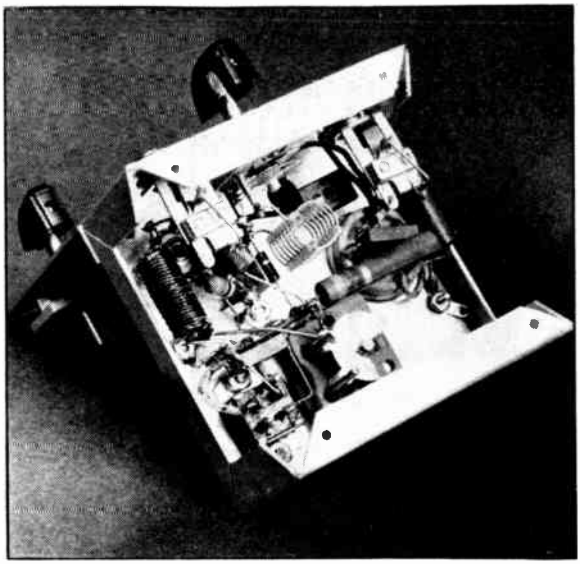


Fig. 13-31 — Bottom view of the 50-Mc. rig. Note the interwound cathode coil at the left.

Provision is made for metering the grid and plate circuits of the final stages by means of jacks in each rig. An approximate check on the final plate currents, sufficient for normal tuning-up purposes, is provided by a 60-ma. pilot lamp connected in the high-voltage lead to the final plate coil. After a few comparisons between the bulb brilliance and observed plate-meter readings it will be possible to estimate the plate current fairly closely by this means. The red jewel in front of the lamp also allows it to serve as a power-on indicator. Off-resonance or no-drive plate current in the 50-Mc. final stage may be sufficient to burn out a 60-ma. pilot lamp, so a 150-ma. bulb may be used during the initial-test phases. Once the rig is adjusted there is little likelihood that the current will exceed 80 ma. or so, which the 60-ma. lamp will take in stride.

The Modulator and Control Circuits

The modulator, Figs. 13-35 and 13-36, is also the power-distribution unit. Control of the power system is by the push-to-talk microphone button, or the toggle switch, S_1, by which the transmitter may be turned on and off conveniently from the test position. This switch is, of course, normally open. The only other control switch is one to be mounted at the operating position to select the band to be used. If only one r.f. section is constructed this remote selector switch (not shown in the schematic diagrams) and its associated power socket, J_2 in Fig. 13-36, can be dispensed with.

The male power plug, P_1 in Fig. 13-36, and the three female power sockets, J_2, J_3 and J_4, are mounted along the back of the modulator chassis. Power details of a typical installation are shown at A and B in this diagram. A 3-wire

Fig. 13-32 — Rear view of the 144-Mc. mobile unit. The copper fin at the side of the final tube is a neutralizing adjustment.

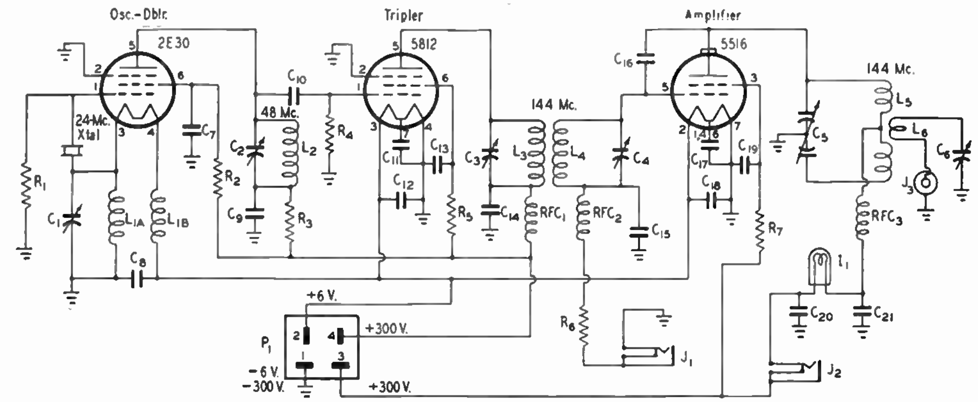


Fig. 13-33 — Schematic diagram of the 144-Mc. r.f. section.

C_1 — 50-$\mu\mu$fd. variable (Millen 20050).
C_2, C_3, C_4 — 15-$\mu\mu$fd. variable (Millen 20015).
C_5 — 6-$\mu\mu$fd.-per-section butterfly variable (Cardwell ER-6-BFS).
C_6 — 35-$\mu\mu$fd. variable (Millen 20035).
C_7, C_8, C_9, C_{11}, C_{12}, C_{13}, C_{14}, C_{15}, C_{17}, C_{18}, C_{19}, C_{20}, C_{21} — 470-$\mu\mu$fd. mica.
C_{10} — 47-$\mu\mu$fd. mica.
C_{16} — Neutralizing-capacitor plate — see text and Fig. 13-32.
R_1, R_4 — 0.1 megohm, ½ watt.
R_2 — 82,000 ohms, ½ watt.
R_3 — 100 ohms, ½ watt.
R_5 — 33,000 ohms, ½ watt.
R_6 — 15,000 ohms, ½ watt.
R_7 — 22,000 ohms, 1 watt.
L_{1A}, L_{1B} — Interwound coils, each 13 turns No. 18 enamel, ⅜-inch diameter.
L_2 — 7 turns No. 18 tinned, ½-inch diameter, ⅞ inch long (B & W Miniductor No. 3002).
L_3, L_4 — Hairpin loops No. 14 wire, 1¼ inches long, ⅞ inch wide. (See bottom view, Fig. 13-34).
L_5 — 6 turns No. 14, c.t., with ⅜-inch space at center, 1½-inch diameter, 1 inch total length.
L_6 — 1¼ turns No. 14 enamel, ⅜-inch diameter.
I_1 — Pilot-lamp assembly with 60-ma. bulb.
J_1, J_2 — Closed-circuit jack.
J_3 — Coaxial output fitting.
P_1 — 4-prong male plug (Jones L-304-AB).
RFC_1, RFC_2, RFC_3 — 1.8-μh. r.f. choke (Ohmite Z-144).

shielded cable can be used between the power sources, B, and the power plug, P_1, on the modulator. The wires carrying the filament current and the generator starting current should, of course, be heavy conductors. The cable shield can be used for the common ground, Pin 2 on P_1.

If the filament selector switch is located at a distance from the modulator the leads from it to J_2 should be of wire capable of carrying 2 amperes without appreciable drop. As indicated in the diagram, there should be 4-conductor cables from J_3 to the 50-Mc. r.f. section, and from J_4 to the 144-Mc. unit.

The modulator uses a single stage, without a speech amplifier. Though this necessitates close talking it makes for economy and simplifies bias problems. It also keeps down power-supply noise (electrical) and car noise (mechanical). With a 300-volt supply there is adequate audio for modulating the final stage of either rig. Bias is supplied by a 30-volt hear-

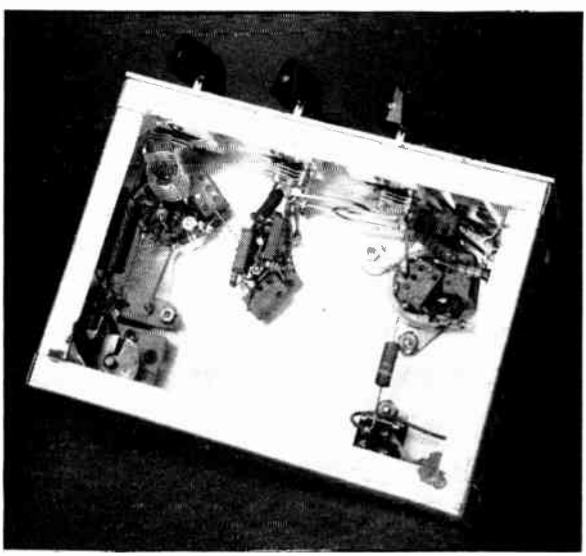


Fig. 13-34 — Bottom view of the 144-Mc. transmitter. Note the hairpin loops in the tripler-plate and amplifier-grid circuits. Oscillator components are at the left, the tripler in the middle, and the amplifier at the right.

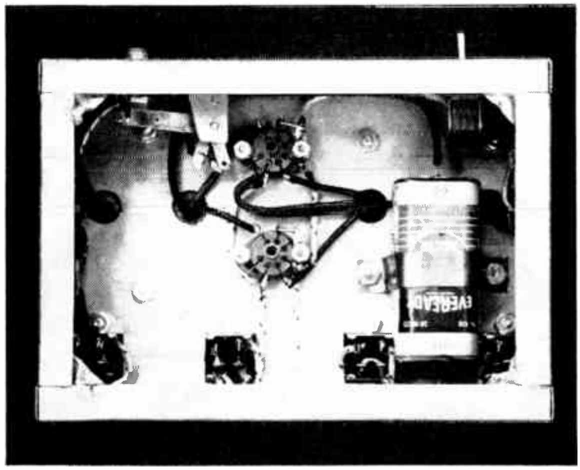


Fig. 13-35 — Bottom view of the modulator and power-distribution unit.

ing-aid battery, which should be good for two years or more of ordinary use.

Testing

Operation of this equipment is similar to that of any transmitter using tetrode tubes, except for the removal of filament voltage during stand-by periods. A supply voltage of 300 is recommended, though lower or higher voltages may be used with suitable modification of the circuit values. No more than 300 volts should be applied to any of the smaller tubes, in any case, and the generator type of supply is recommended.

Bench testing can be done with an a.c. supply, though there will be some hum in the modulation. Operation should be checked, starting with the oscillator, with plate voltage applied to this stage only until it is running properly. An insulated rod, or an empty 'phone plug, can be inserted in the amplifier plate jack to permit tuning the exciter portion without damaging the final tube. The accompanying Table 13-II shows the approximate voltages and currents that will result from use of a 300-volt supply, when the rigs are properly tuned. All controls except the final plate and antenna coupling should be adjusted for maximum final grid current.

The antenna coupling circuit shown will permit the use of almost any coaxial-line-fed antenna system. The proper method of adjustment is to set the coupling at the loosest value that will permit the proper plate current to be drawn when the series condenser is tuned for plate current peak. If the system is properly tuned there will be little, if any, change in the position of the final plate tuning for minimum plate current, with and without the antenna connected to the coaxial output fitting.

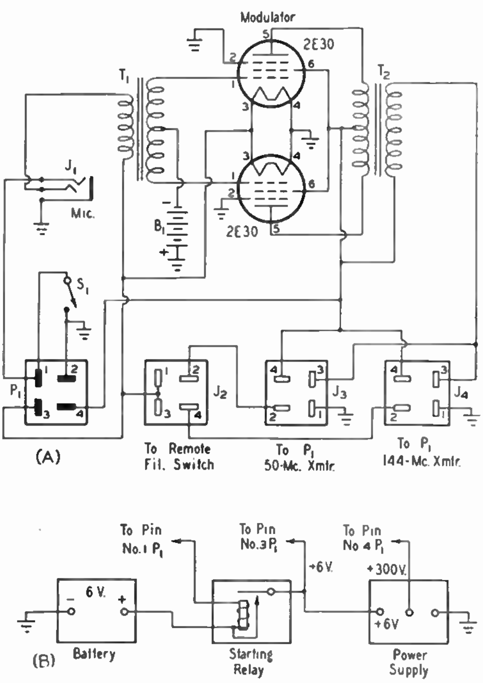


Fig. 13-36 — Schematic diagram of the modulator unit. Chassis size, 2 by 5 by 7 inches. Connections to the power plug and jacks on the unit are shown at A. External power circuits are given in B.

B_1 — Bias battery, 30 volts (Eveready No. 430 hearing-aid type).
J_1 — Microphone jack, double-button type.
J_2, J_3, J_4 — 4-prong female plug (Jones S-304-AB).
P_1 — 4-prong male plug (Jones P-304-AB).
S_1 — S.p.s.t. toggle switch.
T_1 — Microphone transformer (Thordarson T-20A02).
T_2 — Modulation transformer (Stancor A-3845).

Conclusion

Because the form factor of the mobile installation will be different with almost every car, no particular case or mounting is shown. The designs merely show practical parts arrangements and electrical values, leaving the shape and placement of the units to the individual constructor.

A Low-Powered Station for 50 and 144 Mc.

The two small transmitters shown in Fig. 13-37 were designed primarily for use together in mobile service on 50 and 144 Mc., but they may be used as a low-powered two-band home-station, or they may be built and operated separately, if only one of the bands is to be employed. The larger of the two is for 144 Mc., and this unit includes the modulator, though that part of the rig can very well be incorporated in the 50-Mc. unit, if that transmitter is to be used alone. When the two units are connected to a common power source, either one may be used by manipulation of the toggle switches, which apply the heater voltage to the desired circuits.

The r.f. sections are nearly identical, except for the inclusion of a 7F8 tripler stage between the oscillator and the final in the 2-meter unit. Both use Tri-tet oscillators with 6V6GT tubes and fixed-tuned cathode and plate circuits. Harmonic-type crystals are used, 24 to 24.66 Mc. for the 2-meter rig and 25 to 27 Mc. for the 6-meter job, the oscillator doubling in each case. The final stage in both units is an 832 amplifier, the only difference in the circuits being a small amount of neutralization required in the 2-meter rig.

When the two units are used together, 144-Mc. operation requires that switches S_1 and S_2 (Fig. 13-38) be closed, and S_1 in the 50-Mc. unit, Fig. 13-39, left open. For 50-Mc. operation, S_2 is opened, cutting off the r.f. heaters in the 144-Mc. unit, and S_1 in both units is closed. The terminal strips on the backs of the two units are connected in parallel, applying the plate voltages to both at all times, and the heaters of the desired circuits are energized by means of the toggle switches. Switching of the plate voltage is not necessary.

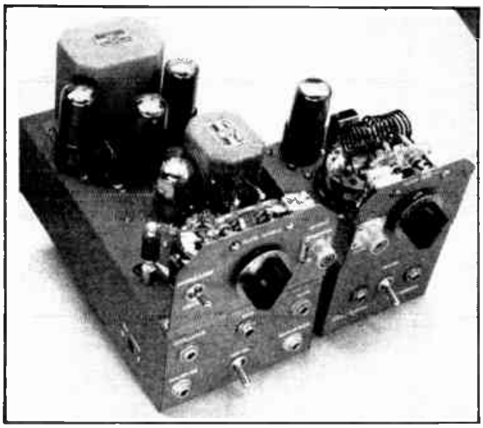


Fig. 13-37— A 2-band set-up for mobile or low-powered fixed-station operation on 50 and 144 Mc. At the left is the 2-meter unit, complete with modulator. The smaller is the 50-Mc. r.f. section. Toggle switches permit use of the modulator with either r.f. section.

THE 144-MC. SECTION

The 144-Mc. unit, Figs. 13-38 and 13-40, includes the modulator and is designed to operate at about 15 watts input with a 300-volt power supply. Meter jacks are provided for measuring the cathode currents of all stages and the grid current of the final. The plate circuits of the oscillator and tripler stages are self-resonant, and are inductively coupled to their following grid circuits.

A small amount of neutralization was required to assure completely-stable operation of the final. The neutralizing condensers, C_{11} and C_{12} in the circuit diagram, are pieces of No. 12 wire extending from the grid of one section of the 832A to the vicinity of the plate of the other section. The wires are crossed at the bottom of the tube socket and go through Millen 32150 bushings mounted in the chassis between the 7F8 and the 832A sockets. It is possible that use of a shielded tube socket would eliminate the tendency toward oscillation in the 832A.

A series-tuned antenna circuit, consisting of C_4 and L_7, is intended for use with any of the low-impedance antenna feed systems commonly used for mobile work. The amount of loading is adjusted by varying the position of the pick-up link, L_7.

The modulator employs a pair of 6V6 or 6V6GT tubes working Class AB. A speech-amplifier stage is not required so long as a single-button carbon microphone is used. Voltage for the microphone is taken from the junction of the two cathode-biasing resistors, R_7 and R_8, thus eliminating the need for a microphone battery.

The microphone and modulation transformers used are both large and expensive for the job at hand and were used only because they happened to be available. The microphone transformer can be any single-button-microphone-to-push-pull-grids transformer and the modulation transformer need not be rated at more than 10 watts. It should be capable of matching a pair of 6V6 tubes to an r.f. load of 5000 to 7000 ohms, depending upon the input at which the 832A is operated.

The photographs of the transmitter show how the parts are mounted on a metal chassis measuring 3 by 5 by 10 inches. The front panel measures 3 by 5 inches and has a ½-inch lip for fastening to the chassis. The construction of the antenna assembly and the method of mounting the components on the panel are identical to the 50-Mc. transmitter. A recommended system of mounting the 832A tube socket is also detailed in the text referring to the 50-Mc. unit.

No special care need be given to the wiring of the audio circuit, but the r.f. leads should be kept as short as possible. The use of four

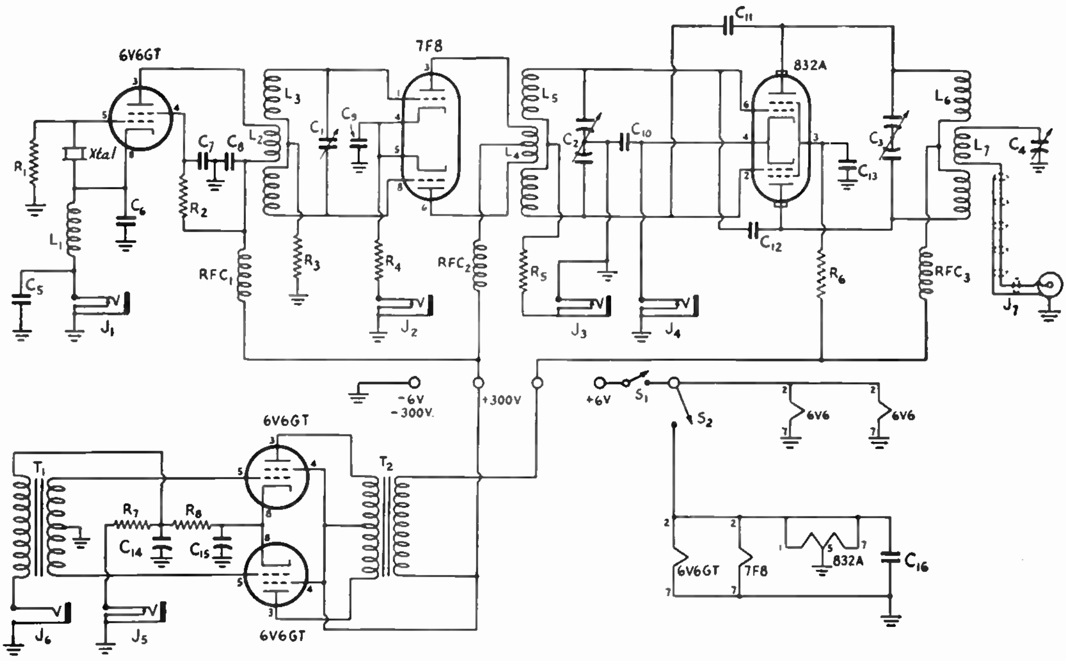


Fig. 13-38 — Circuit diagram of the 144-Mc. r.f. section and modulator.

C_1, C_4 — 3-30-μμfd. mica trimmer.
C_2 — 15-μμfd.-per-section split stator (Bud LC-1660).
C_3 — "Butterfly" condenser, 6 μμfd. per section (Cardwell ER-6-BF/S).
C_5, C_7, C_8 — 0.0047-μfd. mica.
C_6 — 100-μμfd. midget mica.
C_9, C_{10}, C_{13}, C_{16} — 470-μμfd. midget mica.
C_{11}, C_{12} — Neutralizing wires. (See text.)
C_{14}, C_{15} — 10-μfd. 25-volt electrolytic.
R_1 — 0.1 megohm, ½ watt.
R_2 — 47,000 ohms, ½ watt.
R_3 — 33,000 ohms, ½ watt.
R_4 — 470 ohms, ½ watt.
R_5 — 22,000 ohms, ½ watt.
R_6 — 25,000 ohms, 10 watts.
R_7 — 100 ohms, 1 watt.
R_8 — 150 ohms, 1 watt.
L_1 — 3 turns No. 18 enam., close-wound, ½-inch diam.
L_2 — 4 turns No. 18 enam., ⅜ inch long.
L_3 — 10 turns No. 18 enam.; coil wound in two sections with 5 turns each side of L_2, each section ⅜ inch long. A ½ inch is left between windings.
Form for L_2L_3 is a Millen 30003 Quartz-Q stand-off insulator, ¾-inch diam.
L_4 — 3 turns No. 18 enam., ½ inch long, 9⁄16-inch diameter.
L_5 — 2 turns No. 18 enam., interwound with turns of L_4. L_4 and L_5 are wound on a National PRE-3 coil form.
L_6 — 4 turns No. 12 enam., ½-inch i.d., wound in two sections with 2 turns each side of center-tap and a ½-inch space at the center, turns spaced wire diameter.
L_7 — 3 turns No. 12 enam., ½-inch diam., turns spaced wire diameter.
J_1–J_5 — Closed-circuit jack.
J_6 — Open-circuit jack.
J_7 — Coaxial-cable connector.
RFC_1, RFC_2 — 300-μh. r.f. choke (Millen 34300).
RFC_3 — 2.5-mh. r.f. choke (Millen 34102).
S_1, S_2 — S.p.s.t. toggle.
T_1 — Single-button microphone transformer (UTC S-7).
T_2 — Modulation transformer (UTC S-19).

tie-point strips will simplify the mounting and wiring of parts. A single tie point is mounted to the rear of the oscillator tube socket and is used as the junction of R_7, R_8, C_{14} and the primary lead of the microphone transformer. A double tie-point strip is mounted to the right of the crystal socket (as seen in Fig. 13-40). One lug is used as the connecting point for the positive high-voltage lead and the bottom ends of RFC_1 and RFC_2, the bottom of L_1 and the top ends of C_5 and J_1 are connected to the second terminal. The cathode end of L_1 is connected to the cathode side of the crystal socket. The third tie-point strip is mounted on the 832A tube socket and serves as the connecting point between R_4 and J_2; the bottom end of R_6 connects to the high-voltage lead at the second lug. The fourth strip (single lug) is mounted on the frame of C_2 and the leads between R_5 and J_3 join at this point.

The construction of the driver-stage coils is not difficult if the coil forms are properly prepared in advance. A study of Fig. 13-40 will show how the windings are placed on the forms, and the lengths of the windings are given in the parts list. The forms should be marked and drilled to accommodate the windings with the holes for the ends of the windings passing directly through the forms. L_3 should be wound in two sections with the inside ends being soldered together after the winding of L_2 has been completed. The center-taps for L_4 and L_5 are made by cleaning and twisting the wire at the center of each winding. Condenser C_1 is soldered across the grid ends of L_3 before the coil is connected to the tube socket.

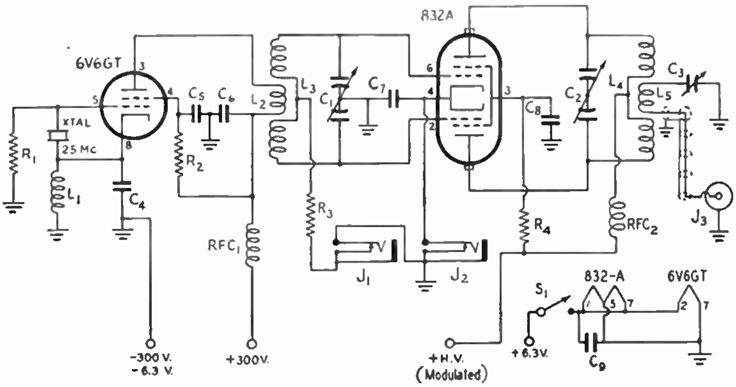


Fig. 13-39 — Circuit diagram of the 6-meter r.f. section.

C_1 — 15-μμfd. per section (Bud LC-1660).
C_2 — "Butterfly" condenser, 15 μμfd. total (Cardwell ER-15-BF/S).
C_3 — 3–30-μμfd. mica trimmer.
C_4 — 100-μμfd. midget mica.
C_5, C_6 — 0.0047-μfd. mica.
C_7, C_9 — 470-μμfd. midget mica.
C_8 — 0.001-μfd. mica.
R_1 — 0.12 meghom, ½ watt.
R_2 — 47,000 ohms, ½ watt.
R_3 — 22,000 ohms, ½ watt.
R_4 — 25,000 ohms, 10 watts.
L_1 — 3 turns No. 18 enameled wire, close-wound, ½-inch diam.
L_2 — 5 turns.
L_3 — 9 turns, 4½ each side of center, with a ⅞-inch space between sections.
L_4 — 10 turns, 5 each side of center, with a ¾-inch space between sections.
L_5 — 3 turns. L_2 through L_5 have an inside diameter of ¾ inch; No. 12 enameled wire, turns spaced wire diameter.
J_1, J_2 — Midget closed-circuit jack.
J_3 — Coaxial-cable connector.
RFC_1 — 10-μh. r.f. choke (Millen 34300).
RFC_2 — 2.5-mh. r.f. choke (Millen 34102).
S_1 — S.p.s.t. toggle switch.

Adjustment and Testing

When testing the transmitter, it is advisable to start with the high voltage applied to the first two stages only. With a 100-ma. meter plugged in J_1 the oscillator cathode current at resonance should be approximately 30 ma. A low-range milliammeter should now be plugged in J_3 and the final grid circuit should be brought into resonance by adjustment of C_2. Proper operation of the tripler stage will be indicated by a cathode current of approximately 20 ma. and a final-amplifier grid current of 2.5 to 3 ma. The tripler grid condenser, C_1, should be retuned after the amplifier grid circuit has been peaked, to assure maximum overall operating efficiency.

The amplifier should be tested for neutralizing requirements after adequate grid drive has been obtained. If a well-shielded tube socket has been used, it is possible that the amplifier grid current will not be affected by tuning the 832A plate circuit through resonance. However, if the grid current does kick down as the plate circuit is tuned, it will be necessary to add the neutralizing wires referred to in the text and parts list as C_{11} and C_{12}. After installation these wires should be adjusted until no kick in grid current is seen as the 832-A plate circuit is tuned through resonance.

Plate and screen voltages can now be applied to the 832A and the plate circuit tuned to resonance, as indicated by a dip in the cathode current to 40 ma. or less. Then a dummy load (a 15-watt light bulb will do) is connected to the antenna jack and the loading adjusted by varying the position of L_7 and the capacitance of C_4, to cause a cathode current of 60 to 70 ma. Approximately 10 ma. of the total cathode current will be drawn by the screen of the 832A and this value should be subtracted from the cathode current in determining the plate input. Amplifier grid-current should be 1.5 to 2 ma. under load.

Modulator cathode current should be 75 ma.; 85 ma. with modulation. The reading will decrease slightly when the microphone is plugged into the circuit. This is caused by the parallel current path that exists when the microphone circuit is completed.

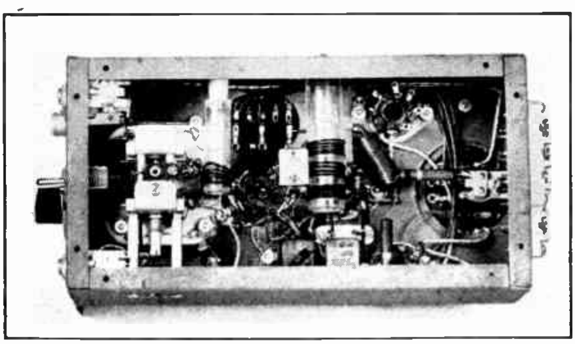


Fig. 13-40 — Bottom view of the 144-Mc. transmitter. The coil forms for L_2L_3 and L_4L_5 are mounted on the side wall of the chassis: the form for L_4L_5 is mounted on a small stand-off insulator so that the windings can be brought out to the center line of the chassis. C_1, the grid condenser for the frequency multiplier, is soldered across the grid ends of L_3. The amplifier grid tuning condenser, C_2, is mounted on metal pillars having a length of 1⅝ inches.

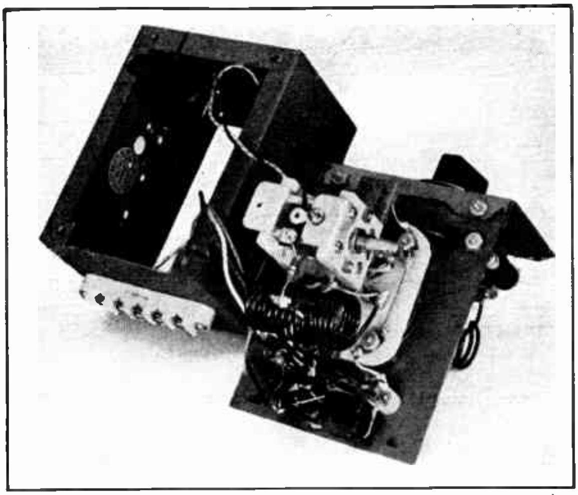


Fig. 13-41 — Bottom view of the mobile transmitter, showing all major components attached to the top plate.

THE 50-MC. PORTION

The 50-Mc. unit, shown in Figs. 13-39 and 13-41, is very similar to the 144-Mc. portion, but for the elimination of the tripler stage. Because of the somewhat lighter load on the power supply, slightly higher power can be run on 50 Mc. In addition, the amplifier operates more efficiently at the lower frequency, permitting inputs up to 30 watts or so if the power is available. Neutralization is not generally required in 50-Mc. operation, but this may not hold true for all physical layouts.

Jacks are provided for measuring the grid and cathode currents of the final stage, and the cathode jack may be used for keying, if c.w. operation is desired. Interstage and antenna coupling circuits are similar to the 144-Mc. section.

The photographs show how a metal box measuring 3 by 4 by 5 inches serves as the chassis for the transmitter. The bottom plate of the box is removed and used as a panel, and is held in place by the screws and nuts that hold the top cover and the box together. In Fig. 13-37 the condenser, C_2, and the antenna jack may be seen mounted on the panel. Metal pillars, ¼ inch long, are used to space the condenser away from the panel. A National FWB polystyrene insulator is used as a mounting support for the antenna coil, L_5, and the insulator is mounted on ¾-inch metal posts. C_3 is supported by its own mounting tabs, and is connected between one end of the pick-up link and ground.

The rear and bottom views of the transmitter show how the rest of the components are laid out on the top plate of the metal box. This plate should be removed from the box while the construction and wiring are being carried on. All of the wiring, with the exception of the d.c. leads to the metering jacks and the input terminals, can be completed in convenient fashion before the top plate is attached to the metal box.

The socket for the amplifier tube is centered on the chassis plate at a point 2⅜ inches in from the front edge, and is mounted below the plate on metal pillars ⅝ inch long. A clearance hole for the 832A, 2¼ inches in diameter, is directly above the tube socket. Sockets for the oscillator tube and the crystal are mounted toward the rear of the chassis.

The oscillator coil, L_2, is mounted on the 6V6 socket; the spare pin, No. 6, of the socket being used as the tie point for the cold end of the plate coil and the other connections that must be made at this part of the circuit. The oscillator cathode coil is mounted between the cathode pin of the 6V6 and a soldering lug placed under the mounting screw of the crystal socket C_5 and C_6 can be seen to the rear of the crystal socket, and RFC_1 is mounted between the tube socket and a bakelite tie-point strip located at the left of the chassis.

The method employed to assure good r.f. grounding of the amplifier components is visible in Fig. 13-41. Soldering lugs are placed beneath the mounting nuts of the 832A socket, and these lugs are joined together with a No. 12 lead which, in turn, is carried on to the common ground point for the oscillator circuit. The filament, cathode, and screen by-pass condensers for the amplifier are all returned to the common ground. These three condensers, C_7, C_8 and C_9, all rest on the 832A tube socket.

The amplifier grid coil, L_3, is self-supporting, with the ends connected to the grid pins of the 832A socket. The tuning condenser, C_1, is actually supported on metal pillars at the right-hand side of the metal box, but the condenser can be wired in place if the operation is carried out in the proper order. First, mount the chassis plate on the box and locate the proper place for the condenser. Next, determine the length of the leads to connect the condenser to the tube socket, and then remove the chassis from the case. The condenser may now be wired into the circuit, and the rigid mounting of C_1, by means of metal posts 1¼ inches long, can be done during the final assembly of the unit.

The grid leak, R_3, is connected between the center-tap of L_3 and a tie-point strip that is mounted on the condenser frame. RFC_2 is mounted toward the front of the chassis, and the grommet-fitted hole to the left of the choke (Fig. 13-41) carries the lead between the plate-voltage terminal and the choke.

The metering jacks and the power terminal strip may now be mounted on the front and rear walls of the metal box. Holes to permit mounting and adjustment of C_1 should also be drilled at this time. Portions of top flanges of the metal case must be cut away in order to provide clearance for the oscillator section and the mounting nut for the amplifier plate choke. After the case, chassis and panel have been fastened together, the wiring of the amplifier plate circuit may be completed.

Test Procedure

A power supply capable of delivering 300 volts at 100 ma. and 6.3 volts at 2 amp. may be used for testing the transmitter. The high voltage should not be applied to the 832A plates until the oscillator has been checked. For initial tests the input voltage can be reduced to approximately 150 volts while the circuits are checked for resonance and proper operation. Squeezing or spreading the turns of the coils should bring the circuits into resonance, as indicated by maximum grid current to the 832A. The grid current should fall to zero, and the plate current of the oscillator tube should rise considerably when the crystal is removed from the socket.

The amplifier plate and screen voltage can be applied at this point. The unloaded cathode current of the amplifier should be about 15 ma., rising to a maximum of 75 or 80 ma. under load, which may be a 15-watt light bulb connected to the antenna jack. C_3 should be adjusted along with the coupling between L_4 and L_5 until maximum output is obtained. The correct degree of loading has been obtained when the plate current at resonance is 10 to 15 ma. below the off-resonance value. The plate tuning condenser, C_2, should be reset each time that a loading adjustment is made.

A final check of voltages and currents should show the following: oscillator and amplifier plate, 300 volts; oscillator screen, 200 volts; amplifier screen, 150 volts; amplifier bias (read at the grid-coil center-tap with a high-resistance voltmeter), 65 volts, negative.

The oscillator plate current should be 28 to 30 ma. and amplifier grid current should be about 3 ma. Under load, the amplifier cathode current should be approximately 60 ma. with 8 or 10 ma. of this amount being drawn by the 832A screen.

Modulation can be supplied by the audio system used in the 2-meter rig shown in Fig. 13-38, or a similar unit may be added, if only 50-Mc. operation is desired.

Transceivers

The transceiver is a combination transmitter-receiver in which, by suitable switching of d.c. and audio circuits, the same tube and r.f. circuit functions either as a modulated transmitting oscillator or as a superregenerative detector. This makes for extreme compactness and light weight, making the transceiver popular for hand-carried portable equipment. It is a compromise with respect to other features, however. The transceiver can be a source of serious interference, and its efficiency is not equal to that of other types of gear wherein separate tubes and circuits are used for transmission and reception.

As a matter of good amateur practice the use of transceivers should be confined to very low-power operation — as in "walkie-talkie" or "handie-talkie" equipment — in the 144-Mc. band, and to experimental low-power operation in the higher-frequency bands. The use of transceivers should be avoided entirely for regular operation on the 144-Mc. band.

CHAPTER 14

V.H.F. Antennas

While the basic principles of antenna operation are essentially the same for all frequencies, certain factors peculiar to v.h.f. work call for changes in antenna technique for the frequencies above 50 megacycles. Here the physical size of multielement arrays is reduced to the point where an antenna system having some gain over a simple dipole is possible in nearly every location, and experimentation with various types of arrays is an important part of the program of most progressive amateurs. The importance of high-gain antennas in v.h.f. work cannot be overemphasized. A good antenna system is often the sole difference between routine operation and outstanding success in this field. By no other means can so large a return be obtained from a small investment as results from the erection of a good directional array.

Design Factors

Beginning with the 50-Mc. band, the frequency range over which antenna arrays should operate effectively is often wider in percentage than that required of lower-frequency systems; thus greater attention must be paid to designing arrays for maximum frequency response, possibly to the extent of sacrificing other factors such as high front-to-back ratio.

As the frequency of operation is increased, losses in the transmission line rise sharply; hence it becomes more important that the line be matched to the antenna system correctly. Because any v.h.f. transmission line is long, in terms of wavelength, it is often more effective to use a high-gain array at relatively low height, rather than to employ a low-gain system at great height above ground, particularly if the antenna location is not completely shielded by heavy foliage, buildings, or other obstructions in the *immediate* vicinity.

This concept is in direct contrast to early notions of what was most desirable in a v.h.f. antenna system. An appreciable clearance above surrounding terrain is desirable, but great height is by no means so all-important as it was once thought to be. Outstanding results have been obtained by many v.h.f. workers, especially on 50 and 144 Mc., with antennas not more than 25 to 40 feet above ground. DX can be worked on 50 Mc. with arrays as low as a half-wave above the ground level.

Polarization

Practically all the early work on frequencies above 30 Mc. was done with vertical antennas, probably because of the somewhat stronger field in the immediate vicinity of a vertical system. When v.h.f. work was confined to almost pure line-of-sight distances, the vertical dipole produced a stronger signal at the edge of the working range than did the same antenna turned over to a horizontal position. With the advent of high-gain antennas and extended operating ranges, horizontal systems began to assume importance in v.h.f. work, especially in parts of the country where a considerable degree of activity had not already been established with verticals.

Numerous tests have shown that there is very little difference in the effective working range with either polarization, *if* the most effective element arrangements are used and the same polarization is employed at both ends of the path. Vertical polarization still has its adherents among 50-Mc. enthusiasts and much fine work has been done with vertical antennas, but an effective horizontal array is somewhat easier to build and rotate. Simple 2-, 3- or 4-element horizontal arrays have proven extremely effective in 50-Mc. work, and the postwar era has seen an increase in the use of such arrays which has amounted to standardization on horizontal polarization.

The picture is somewhat different when one goes to 144 Mc. and higher. At these frequencies, the most effective vertical systems (those having two or more half-wave elements, vertically stacked) are more easily erected than on 50 Mc. Important, in considering the polarization question, is the existence of numerous 144-Mc. mobile stations whose antenna systems must, of necessity, be vertical. While horizontal polarization will undoubtedly find increased favor at 144 Mc. and higher, particularly for point-to-point work in rural areas, it is probable that vertical polarization will continue in use for some years to come, particularly in areas where activity has been established with vertical systems. Under certain conditions, notably a station directly in the shadow of a hill, there may be a considerable degree of polarization shift, but ordinarily it may be assumed that best results in 144-Mc. work will be obtained by matching the antenna polarization of the stations one desires to contact.

Impedance Matching

Because line losses tend to be much higher in v.h.f. antenna systems, it becomes increasingly important that feedlines be made as nearly "flat" as possible. Transmission lines commonly used in v.h.f. work include the open-wire line of 500 to 600 ohms impedance, usually spaced about two inches; the polyethylene-insulated flexible lines, available in impedances of 300, 150, 100 and 72 ohms; and coaxial lines of 50 to 90 ohms impedance. These may be matched to dipole or multielement antennas by any of several arrangements detailed below.

The "J"

Used principally as a means of feeding a stationary vertical radiator, around which parasitic elements are rotated, the "J" consists of a half-wave vertical radiator fed by a quarter-wave matching section, as shown at A, Fig. 14-1. The spacing between the two sides of the matching section should be two inches or less, and the point of attachment of the feedline will depend on the impedance of the line used. The feeder should be slid along the matching section until the point is found that gives the best operation. The bottom of the matching section may be grounded for lightning protection. A variation of the "J" for use with coaxial-line feed is shown at B in Fig. 14-1. The "J" is also useful in mobile applications.

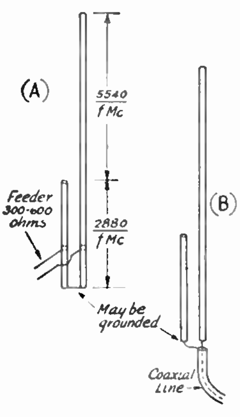


Fig. 14-1 — Two versions of the "J" antenna, often used in mobile installations, or in vertical arrays where parasitic elements may rotate around a fixed radiator.

The Delta or "Y"-Match

Probably the simplest arrangement for feeding a dipole or parasitic array is the familiar delta, or "Y"-match, in which the feeder system is fanned out and attached to the radiator at a point where the impedance along the element is the same as that of the line used. Information on figuring the dimensions of the delta may be found in Chapter Ten. Chief weakness of the delta is the likelihood of radiation from the matching section, which may interfere with the effectiveness of a multielement array. It is also somewhat unstable mechanically, and quite critical in adjustment.

The "Q" Section

An effective arrangement for matching an open-wire line to a dipole, or to the driven element in a 2- or 3-element array having wide (0.25 wavelength or greater) spacing, is the "Q" section (Chapter Ten). This consists of a quarter-wave line, usually of ½-inch or larger tubing, the spacing of which is determined by the impedance at the center of the array. The parallel-pipe "Q" section is not practical for matching multielement arrays to lines of lower impedances than about 600 ohms, nor can it be used effectively with close-spaced parasitic arrays. The impedance of the "Q" section required in these cases is lower than can be obtained with parallel sections of tubing of practical dimensions. A quarter-wave section of coaxial or other low-impedance line is a commonly-used means of matching a line of 300 to 600 ohms impedance to the low center impedance of a 3- or 4-element array. The length of such a line will depend on the velocity of propagation (propagation factor) of the line used. The propagation factors of all the commonly-used lines are given in table form in Chapter Ten.

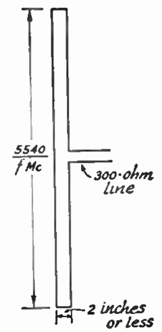


Fig. 14-2 — Details of the folded dipole.

In some installations it may be more convenient to use a line of greater length than a single quarter wave for matching purposes, in which case any odd multiple of a quarter wavelength may be used. The exact length required may be determined experimentally by shorting one end of the line and coupling it to a source of r.f., and trimming the line length until maximum loading is obtained at the center frequency of the operating range.

The "T"-Match

The principal disadvantages of the delta system can be overcome through the use of the arrangement shown in Figs. 14-5 and 14-13, commonly called the "T"-match. It has the advantage of providing a means of adjustment (by sliding the clips along the parallel conductors), yet the radiation from the matching arrangement is lower than with the delta, and its rigid construction is more suitable for rotatable arrays. It may be used with coaxial lines of any impedance, or with the various other forms of transmission lines up to 300 ohms. The position of the clips should, of course, be adjusted for maximum loading and minimum standing-wave ratio, the latter being most important as an indication of proper setting. The "T" system is particularly well suited for use in all-metal "plumbing" arrays.

The Folded Dipole

Probably the most effective means of matching various lines to the wide range of antenna impedances encountered in v.h.f. antenna work is the folded dipole, shown in its simplest form in Fig. 14-2. When all portions of the dipole are of the same conductor size, the impedance at the feed-point is equal to the *square* of the number of elements in the folded dipole times the normal center impedance which would be present if only a conventional split half-wave radiator were used. Thus, the simple folded dipole of Fig. 14-2 has a feed-point impedance of 4 × 72, or approximately 288 ohms. It may be fed with the popular 300-ohm line without appreciable mismatch. If a three-wire dipole were used, the step-up in impedance would be *nine* times. Note that this step-up occurs *only* if all portions of the folded dipole are the same conductor size.

The impedance at the feed-point of a folded dipole may also be raised by making the fed portion of the dipole smaller than the parallel section. Thus, in the 50-Mc. array shown in Fig. 14-4 the relatively low center impedance of a 4-element array is raised to a point where it may be fed directly with 300-ohm line by making the fed portion of the dipole of ¼-inch tubing, and the parallel section of 1-inch. A 3-element array of similar dimensions could be matched by substituting ¾-inch tubing in the unbroken section. Conductor ratios and spacings may be obtained from the folded-antenna nomogram in Chapter Ten.

Antenna Systems for 50 and 144 Mc.

Since the same basic principles apply to all antennas regardless of frequency, little discussion is given here of the various simple dipoles that may be used when nondirectional systems are desired. Details of such antennas may be found in Chapter Ten, and the only modification necessary for adaptation to use on 50 Mc. or higher is the reduction in length necessary for increased conductor diameter at these frequencies.

A Simple 2-Element Array

A simple but effective array which requires no matching arrangement is shown in Fig. 14-3. Its design takes into account the drop in center impedance of a half-wave radiator when a parasitic element is placed a quarter wavelength away. A director element is shown, as the drop in impedance using a slightly-shortened parasitic element is just about right to provide a good match to a 50-ohm coaxial line. The element lengths are not extremely critical in such a simple system, and the figures presented may be used with satisfactory results.

A 4-Element Array

The importance of broad frequency response in any antenna designed for v.h.f. work cannot be overlooked. The disadvantage of all parasitic systems is that they tend to tune quite sharply, and thus are often effective over only a small portion of a given band. One way in which the response of a system can be broadened out is to increase the spacing between the

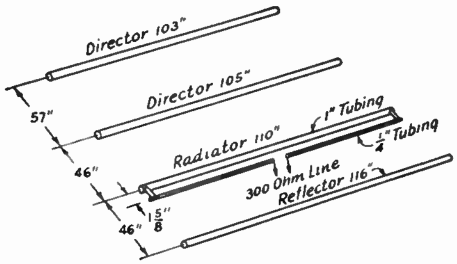


Fig. 14-4 — Dimensional drawing of a 4-element 50-Mc. array. Element length and spacing were derived experimentally for maximum forward gain at 50.5 Mc.

parasitic elements to somewhat more than the 0.1 or 0.15 wavelength normally considered to provide optimum front-to-back ratio. Some broadening may also be obtained by making the directors slightly shorter and the reflector slightly longer than the optimum value. The folded dipole is useful as the radiator in such an array, as its over-all frequency response is somewhat broader than other types of driven elements.

A 4-element array for 50 Mc. having an effective operating range of about 2 Mc. is shown in Figs. 14-4 and 14-5. It employs a folded dipole having nonuniform conductor size. Reflector and first director are spaced 0.2 wavelength from the driven element, while the forward director is spaced 0.25 wavelength. The spacing and element lengths given were derived experimentally, and are those that give optimum forward gain at the expense of some front-to-back ratio. As the latter quality is not of great value in 50-Mc. work, it can be neglected entirely in the tuning procedure for such an array.

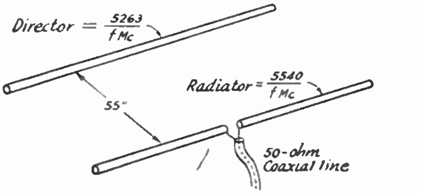


Fig. 14-3 — A simple 2-element array for 50 Mc. No matching devices are needed with this arrangement.

The dimensions given are for peak performance at 50.5 Mc. For other frequencies, the length of the folded dipole in inches should be figured according to the formula

$$L = \frac{5540}{f_{Mc.}}$$

The reflector will be 5 per cent longer, the first director 5 per cent shorter, and the second director 6 per cent shorter than the driven element. A broadening of the response may be obtained, at a slight sacrifice in forward gain, by adding to the reflector length and subtracting from the director lengths. For those interested in experimenting with element lengths,

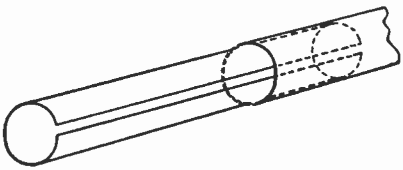


Fig. 14-5 — **Detail drawing of inserts which may be used in the ends of the elements of a parasitic array to permit accurate adjustment of element length.**

slotted extensions may be inserted in the ends of the various elements, other than the dipole, as shown in Fig. 14-5. A 3-element array may be built, using the same general dimensions, except that the unbroken section of the folded dipole, in this case, should have a ¾-inch diameter element in place of the 1-inch tubing used in the 4-element array.

Stacked Antennas

The radiation angle of a 50-Mc. antenna system may be lowered, with a resulting improvement in operating range, by stacking two or more parasitic arrays and feeding them in phase. At spacings of ½ to ⅝ wavelength a gain of 4 db. or more may be realized by the stacking of two arrays. Examples for 50 and 144 Mc. are combined in the dual array shown in Fig. 14-6.

Two 50-Mc. 4-element arrays are mounted one-half wavelength apart, with a similar dual system for 144 Mc. in the middle of the space between them. The structure is all-metal design. Booms for the 50-Mc. portion are 1½-inch 24ST dural tubing 128 inches long. Elements are ¾-inch tubing of the same alloy, forced through holes in the booms. Director spacing is 0.2 wavelength, reflector spacing 0.15 wavelength. The booms are mounted on the vertical member (a 1½-inch o.d. pipe) by means of blocks of wood, the only nonmetallic parts employed. These were made from pieces of two-by-four one foot long. A hole the size of the mast is made in the block near one side, at the middle of the block lengthwise. The block is then sawed lengthwise in a vertical plane, through the middle of this hole. Bolting the two portions together provides a tight fit around the vertical pipe. The boom is bolted to the block at three points. This method of mounting provides a rigid assembly. The booms should be bonded to the main support to provide lightning protection.

Booms in the 2-meter array are of 1-inch tubing, and elements of ⁷⁄₁₆-inch, mounted through the booms as in the larger array. The vertical member is 1½-inch dural tubing, attached to the main pipe with "U" bolts. Element spacing is 0.2 wavelength throughout.

A double version of the "Q" system of matching is used in both arrays. Folded-dipole radiators are used in the 6-meter portion, and "T"-matched dipoles in the 2-meter array, but a similar dipole arrangement could, of course, be used in both.

The main transmission line for each array is 300-ohm Twin-Lead. The method of feed was checked out for minimum standing-wave ratio with one bay alone; then the phasing section for the two bays was proportioned so that it would serve as a "Q" section as well. Dimensions for both arrays are given in Table 14-I. The feedlines are brought at right angles from

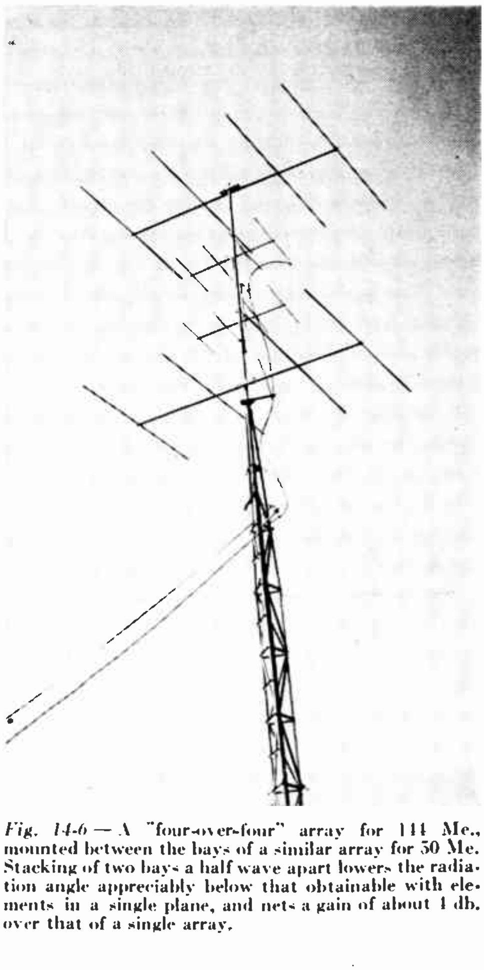


Fig. 14-6 — **A "four-over-four" array for 144 Mc., mounted between the bays of a similar array for 50 Mc. Stacking of two bays a half wave apart lowers the radiation angle appreciably below that obtainable with elements in a single plane, and nets a gain of about 4 db. over that of a single array.**

the phasing sections to stand-off insulators on the main vertical support. They drop vertically to a combination tie point and bearing, just below the lower boom of the 6-meter array. From this anchor, which rotates with the beams, they drop loosely to a fixed tie point, with enough slack left to permit slightly more than 360 degrees of rotation.

The fed sections of the 50-Mc. folded dipoles are made of 3/16-inch copper tubing, mounted on 5/8-inch cone stand-offs. The outer ends are supported on metal pillars of the same length. Two stand-offs are used for each side of the dipole; otherwise the rather soft tubing tends to sag and disturb the spacing between it and the larger element. The copper tubing is flattened in a vise at the points where it is to be mounted. The 4-to-1 conductor ratio, and the spacing of one inch, center to center, between the two conductors gives the necessary impedance step-up to match 300-ohm line, in a 4-element array of the spacings mentioned earlier in this section.

A similar arrangement might have been used in the 2-meter array, but the "T"-match was substituted because a suitable conductor ratio was not so practical with the smaller-sized elements used. Adjusting clips for the "T" section were made from grid clips slipped over the respective elements and soldered together in such a position as to give a spacing of about 1¼ inches, center to center. A one-inch ceramic stand-off was used on each section, to hold the "T" section in alignment with the main element. The phasing section is the same as in the larger array: No. 12 wire spaced one inch. The point of connection between the "T" section and the dipole turned out to be approximately 5 inches from the center, but this should be adjusted for minimum standing-wave ratio.

Phased Arrays

Superior performance is obtainable on 144 Mc. and higher by using curtains of 4, 6, 8 or more driven half-wave elements, arranged in pairs fed in phase, and backed up by reflectors. Figs. 14-7 to 14-9 show 12- and 16-element arrays that are capable of more than 12 and 14 db. gain, respectively. The supporting structures required by such arrays would make them out of the question for lower frequencies, but for 144 Mc. and higher they are relatively easy to build and erect. Their dimensions are not particularly critical, and careful adjustment of the elements is not required for good results. The frequency response of arrays having several driven elements is broader than that of systems in which the gain is built up through the use of additional parasitic elements.

TABLE 14-I

Dimensions of the 6- and 2-Meter Stacked Arrays, in Inches

	Radiator	*Reflector*	*1st Director*	*2nd Director*
50 Mc.	110	116	105	103
144 Mc.	38	40	36	35¾

	Phasing Line	*Reflector Spacing*	*Director Spacing*
50 Mc.	114	34	46
144 Mc.	39½	15¾	15¾

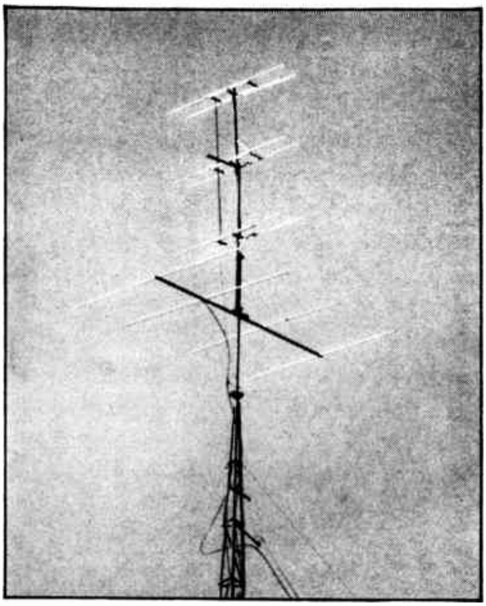


Fig. 14-7 — A 12-element array for 144 Mc. This installation, designed for polarization tests, has its boom mounted on a hinge to permit use of the array in either a horizontal or vertical position. The lower array is a 50-Mc. 4-element beam.

The 12-element array, Figs. 14-7 and 14-8, has a similar pattern in both horizontal and vertical planes. The photograph shows an experimental set-up in which the array was mounted on a door hinge, in order to permit its use in horizontal-vertical tests. The horizontal radiation pattern of the 16-element array is somewhat sharper when it is used in a vertical position, but it is a highly effective antenna either way.

The elements need not be larger than half-inch diameter, and smaller sizes can be used if desired, so the entire structure can be made light in weight and still have considerable strength. The phasing sections may be No. 14 or 16 wire, spaced 1 to 1½ inches. They are transposed in both sides of the 12-element array, and in the two end sections of the 16-element.

Either array may be fed with 300-ohm Twin-Lead, connected as shown in the drawings. The feed impedance of the 12-element array is brought down by spacing the reflectors 0.15 wavelength, making it possible to connect the transmission line to the center pair of elements directly without a matching device. The feed impedance of the 16-element array may be somewhat lower than 300 ohms, but the mismatch is not serious and it may be disregarded if the transmission line is relatively short. If a long line is necessary it may

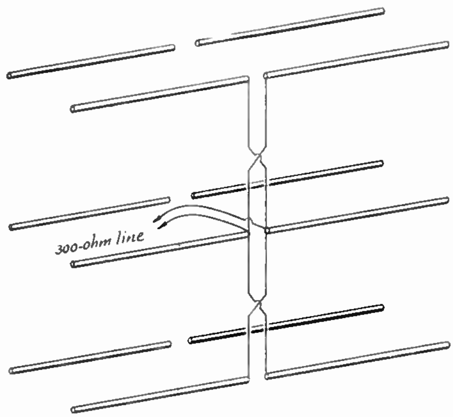


Fig. 14-8 — Element arrangement and feed system of the 12-element array. Reflectors are spaced 0.15 wavelength behind the driven elements.

be desirable to install an adjustable "Q" section at the feed point. This can be made of two 20-inch tubes of the same material as is used for the driven elements, mounted so that the spacing between them can be adjusted for lowest standing-wave ratio. The feed impedance of an array having several driven elements is subject to many variables, making some sort of adjustable impedance-matching device highly desirable if long feedlines are used.

Element lengths and spacings may be taken from Table 14-I, except for the spacing between the driven elements and the reflectors. This should be 12 inches for the 12-element array and 17 inches for the 16-element.

Long-Wire Antennas

Where long-wire systems designed for use on lower frequencies are available they may often be used on the v.h.f. bands with good results, particularly if the feedlines are not too long. "V" and rhombic antenna systems designed expressly for the v.h.f. bands are small enough in size to be used in many locations where similar arrays for lower frequencies would be out of the question. The polarization of long-wire systems is normally horizontal, but in locations where they have a downward slope they may also have a considerable vertical component. Their polarization discrimination is seldom as sharp as that of systems using half-wave elements.

Information on the various types of long-wire arrays will be found in Chapter Ten. At 144 Mc. and higher it is relatively easy to stack two or more "V" or rhombic arrays a half-wave apart. This improves their performance considerably, but makes them essentially one-band devices.

Arrays for 220 and 420 Mc.

The use of a high-gain antenna system is almost a necessity if work is to be done over any great distance on 220 and 420 Mc. Experimentation with antenna arrays for these frequencies is fascinating indeed, as their size is so small as to permit trying various element arrangements and feed systems with ease. Arrays for 420 Mc., particularly, are convenient for use in demonstrations of antenna principles, as even high-gain systems may be of table-top proportions.

Any of the arrays described previously may be used on these bands, but those having a number of driven elements fed in phase will be most desirable. The 12- and 16-element arrays, Figs. 14-8 and 14-9, may be adapted to use on 220 or 420 Mc. by using the dimensions given in Table 14-II.

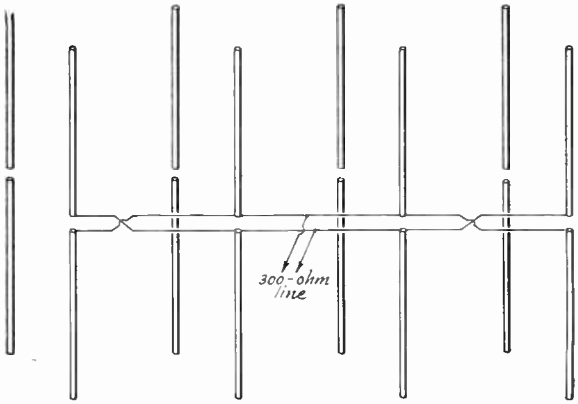


Fig. 14-9 — Schematic drawing of a 16-element array for 144 Mc. or higher bands. A variable "Q" section may be inserted at the feed point if accurate matching is desired.

The use of a plane reflector, in place of the parasitic reflectors used in the 144-Mc. models, is highly desirable when phased systems are used on higher bands. The spacing between the driven elements and the reflecting plane is not particularly critical, except as it affects the feed impedance of the system. Maximum gain occurs in the region around 0.1 to 0.15 wavelength, with the feed impedance being lowest with the closest spacing. The feed impedance is highest at approximately 0.3-wavelength spacing. The reflector has no effect on the feed impedance when a spacing of 0.22 wavelength is used. As the gain is nearly constant from 0.1 to 0.25 wavelength, it may be seen that the spacing may be varied to achieve an impedance match.

An advantage of the plane reflector is that it may be used for two arrays, incorporating horizontal and vertical polarization on opposite sides of the plane, or providing two-band operation, as is done in the array for 220 and 420 Mc. shown in Fig. 14-10. Six driven elements for 220 Mc. are used on one side, arranged in a manner similar to the driven elements in the 12-element array for 144 Mc. described earlier in this chapter. The 420-Mc. side uses 16 driven elements arranged in two sets of 8 each.

These two sets of elements are mounted one above the other with their ends approximately one-half wavelength apart. This dimension is not critical, though maximum gain is obtained with end-to-end spacings of about a half wavelength. The two pairs of phasing wires are connected by means of one-wavelength sections of 300-ohm Twin-Lead at the middle of the array. This junction, which has an impedance of about 150 ohms, is fed with 300-ohm line through an adjustable "Q" section.

The one-wavelength sections of 300-ohm line are 21¾ inches long, this figure taking the propagation factor of the line into account. The "Q" section may be made of the same material as the elements, or any available tubing, from ¼- to ½-inch diameter, may be used. As proper matching is extremely important at 420 Mc. the spacing of this "Q" section should be adjusted carefully for minimum standing-wave ratio.

TABLE 14-II

Element Lengths and Spacings, in Inches, for 12- and 16-Element Arrays for 220 and 420 Mc.

Freq. (Mc.)	*Driven Element*	*Reflector*	*Phasing Section*	*Reflector 12-El.*	*Spacing 16-El*	*"Q" Section*
220	24⅞	26⅛	25⅞	7¾	11½	13
420	12¾	13⅜	13¼	4	5¾	6⅝

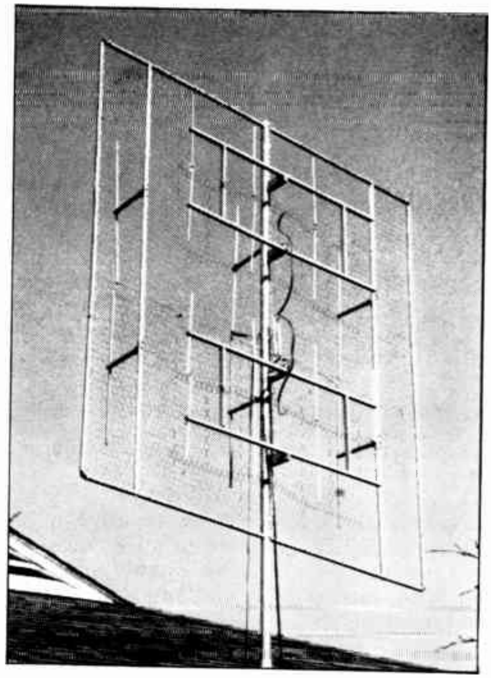


Fig. 14-10 — A two-band screen-reflector array. One side has 16 driven elements for 420 Mc., and the reverse side has 6 half-waves in phase for 220 Mc. Both sets of elements are spaced 0.15 wavelength from the reflecting plane.

The reflecting plane is 6 feet square. This is larger than necessary for the 420-Mc. system, the size being determined by the 220-Mc. side. Chicken wire of 1-inch mesh is used for the screen. Wire netting, sheet metal, or closely-spaced wires may be substituted. The size of the reflector is not critical, except that it should extend at least a quarter wavelength beyond the area covered by the driven elements. A plane-reflector array has slightly more gain than is obtained with the same number of driven elements backed up by parasitic reflectors. The frequency response is wider and it has a considerably higher front-to-back ratio. The principal dimensions may be taken from Table 14-II.

Mobile and Portable Antennas

A common type of antenna employed for mobile operation on 50 and 144 Mc. is the quarter-wave radiator which is fed with a coaxial line. The antenna, which may be a flexible telescoping "fish pole," is mounted in any of several places on the car. The inner conductor of the coaxial line is connected to the antenna, and the outer conductor is grounded to the frame of the car. Quite a good match may be obtained by this method with the 50-ohm coaxial line now available; however, it is well to provide some means of tuning the system, so that all variables can be taken care of. The simplest tuning arrangement consists of a variable condenser connected between the low side of the transmitter coupling coil and ground, as shown in Fig. 14-11. This condenser should have a maximum capacitance of 75 to 100 $\mu\mu$fd. for 50 Mc., and should be adjusted for maximum loading with the least coupling to the transmitter. Some method of varying the coupling to the transmitter should be provided.

The short antenna required for 144 Mc.

(approximately 19 inches) permits mounting the antenna on the top of the car. Such an arrangement provides good coverage in all directions, the car body acting as a ground plane. When the antenna is mounted else-

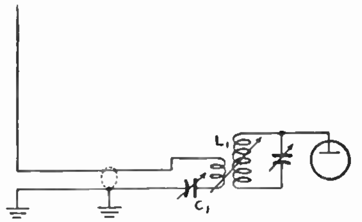


Fig. 14-11 — Method of feeding quarter-wave mobile antennas with coaxial line. C_1 should have a maximum capacitance of 75 to 100 μμfd. for 28- and 50-Mc. work. L_1 is an adjustable link.

where on the car, it is apt to show quite marked directional characteristics. Because of this it is desirable to make provisions for the use of the same antenna for both transmitting and receiving.

A Collapsible Array for 50 Mc.

The best antenna possible for operation under mobile conditions is not particularly effective, as compared with antenna systems normally used in fixed-station work. To make the most of the fine opportunities for DX work afforded by countless high-altitude locations which are accessible by car, it is helpful to have some sort of collapsible antenna array which can be assembled "on the spot." Even a simple array like the one shown in Figs. 14-12 and 14-13 will effect a great improvement in the operating range of the low-powered gear normally used for mobile operation. This one is designed for 50-Mc. use, but similar arrangements can be made for other frequencies.

The array shown is a 2-element system, comprised of a radiator which is fed with coaxial line by means of a "T"-match, and a reflector which is spaced 0.15 wavelength in back of the driven element. It is made entirely of ¾-inch dural tubing, except for the vertical support, which is 1-inch tubing of the same material. A suggested method of mounting is shown in Fig. 14-12. A short length of 1 × 2-inch or larger wood is bolted to the car bumper. A piece of ¾-inch dural tubing is bolted to this upright, and the 1-inch vertical section of the array slips over the top of the ¾-inch section. The array is turned by means of ropes attached to the reflector element. Height of the array may be increased over that shown by using a longer wooden support, in which case it is desirable to use a 2 × 2 for greater strength. An

Fig. 14-12 — A 2-element collapsible array for 50-Mc. portable use.

anchoring pin made from a spike inserted in the bottom end of the wooden support is helpful to prevent tilting of the array. With such a device embedded in the ground, the whole assembly will remain rigid, which is helpful in the high winds usually encountered in mountain-top locations. Portability is provided by making the elements in three sections, with

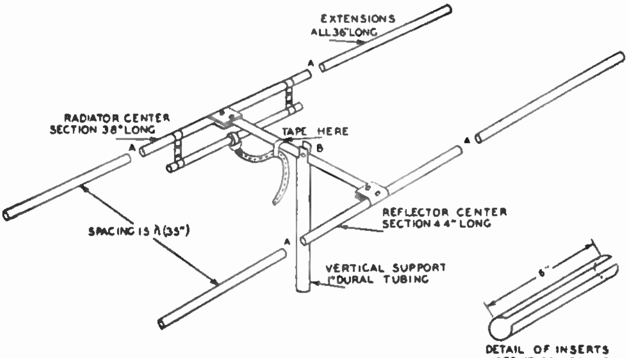


Fig. 14-13 — Detail drawing of the collapsible 50-Mc. array shown in Fig. 14-12. All parts except the vertical support, which is 1 inch in diameter, are made of ¾-inch dural tubing. For carrying purposes, it is taken apart at points *A* and *B*, inserts of slotted dural tubing being used at points *A* to hold the sections together. All extensions are the same length, the difference in element length being provided by the length of the center sections.

the end sections all the same length. The center section of the radiator is 6 inches shorter than that of the reflector.

The fed section of the "T" matching device is composed of two pieces of ¾-inch dural tubing about 14 inches long. The two sections are held together mechanically, but insulated electrically, by a piece of polystyrene rod which is turned down just enough to make a tight fit in the tubing. The inner and outer conductors of the coaxial line are fastened to the two inside ends of the matching section. Clips made of spring bronze are used for connection between the radiator and the "T." The position of these should be adjusted for maximum loading and minimum standing-wave ratio on the line.

This antenna system may be used as a dipole on 29 Mc. by plugging the reflector sections into the driven element, thus bringing its over-all length to approximately that of a half-wave for the high end of the 10-meter band.

Miscellaneous Antenna Systems

Coaxial Antennas

With the "J" antenna radiation from the matching section and the transmission line tends to combine with the radiation from the antenna in such a way as to raise the angle of radiation. At v.h.f. the lowest possible radiation angle is essential, and the coaxial antenna shown in Fig. 14-14 was developed to eliminate feeder radiation. The center conductor of a 70-ohm concentric transmission line is extended one-quarter wave beyond the end of the line, to act as the upper half of a half-wave antenna. The lower half is provided by the quarter-wave sleeve, the upper end of which is connected to the outer conductor of the concentric line. The sleeve acts as a shield about the transmission line and very little current is induced on the outside of the line by the antenna field. The line is nonresonant, since its characteristic impedance is the same as the center impedance of the half-wave antenna. The sleeve may be made of copper or brass tubing of suitable diameter to clear the transmission line. The coaxial antenna is somewhat difficult to construct, but is superior to simpler systems in its performance at low radiation angles.

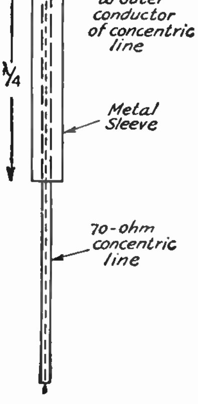


Fig. 14-14 — Coaxial antenna. The insulated inner conductor of the 70-ohm concentric line is connected to the quarter-wave metal rod which forms the upper half of the antenna.

Cylindrical Antennas

Radiators such as are used for television and broad-band FM are of interest in amateur v.h.f. operation because they work at high efficiency without adjustment throughout the width of an amateur band.

At the very-high frequencies an ordinary dipole or equivalent antenna made of small wire is purely resistive only over a very small frequency range. Its Q, and therefore its selectivity, is sufficient to limit its optimum performance to a narrow frequency range, and readjustment of the length or tuning is required for each narrow slice of the spectrum. With tuned transmission lines, the effective length of the antenna can be shifted by retuning the whole system. However, in the case of antennas fed by matched-impedance lines, any appreciable frequency change requires an actual mechanical adjustment of the system. Otherwise, the resulting mismatch with the line will be sufficient to cause significant reduction in power input to the antenna.

A properly designed and constructed wide-band antenna, on the other hand, will exhibit very nearly constant input impedance over several megacycles.

The simplest method of obtaining a broad-band characteristic is the use of what is termed a "cylindrical" antenna. This is no more than a conventional doublet in which large-diameter tubing is used for the elements. The use of a relatively large diameter-to-length ratio lowers the Q of the antenna, thus broadening the resonance characteristic.

As the diameter-to-length ratio is increased, end effects also increase, with the result that the antenna must be made shorter than a thin-wire antenna resonating at the same frequency. The reduction factor may be as much as 20 per cent with the tubing sizes commonly used for amateur antennas at v.h.f.

Cone Antennas

From the cylindrical antenna various specialized forms of broadly-resonant radiators have been evolved, including the ellipsoid,

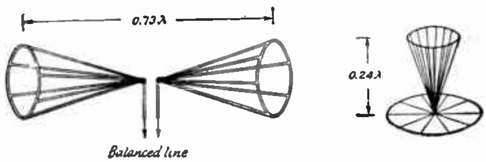


Fig. 14-15 — Conical broad-band antennas have relatively constant impedance over a wide frequency range. The three-quarter wavelength dipole at left and the quarter-wave vertical with ground plane at right have the same input impedance — approximately 65 ohms. Sheet-metal or spine-type construction may be used.

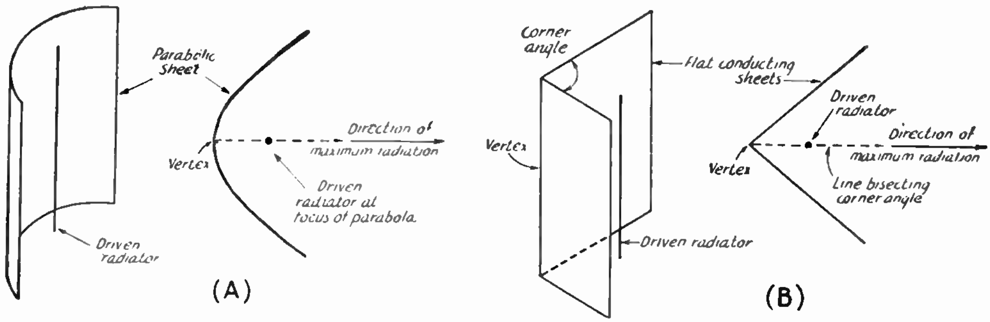


Fig. 14-16 — Plane sheet reflectors for v.h.f. and u.h.f. A shows a parabolic sheet and B a square-corner reflector.

spheroid, cone, diamond and double diamond. Of these, the conical antenna is perhaps the most interesting. With large angles of revolution the characteristic impedance can be reduced to a very low value suitable for extremely wide-band operation. The cone may be made up either of sheet metal or of multiple wire spines, as in Fig. 14-15.

Plane Sheet Reflectors

The small physical size of v.h.f. antennas makes practical many methods not feasible on lower frequencies. For example, a plane flat-sheet reflector may be used with a half-wave dipole, obtaining gains of 5 to 7 db. Much higher gains are attainable with a number of stacked dipoles fed in phase, mounted in front of a reflecting plane. Such an arrangement is called a "billboard" array.

Plane reflectors need not be constructed of solid sheets. Wire mesh, or a grid of closely-spaced parallel-wire spines, is more easily erected and offers lower wind resistance.

Parabolic Reflectors

A plane sheet may be formed into the shape of a parabolic curve and used with a driven radiator situated at its focus, to provide a highly-directive antenna system. If the parabolic reflector is sufficiently large so that the distance to the focal point is a number of wavelengths, optical conditions are approached and the wave across the mouth of the reflector is a plane wave. However, if the reflector is of the same order of dimensions as the operating wavelength, or less, the driven radiator is appreciably coupled to the reflecting sheet and minor lobes occur in the pattern. With an aperture of the order of 10 or 20 wavelengths, a beam-width of approximately 5 degrees may be achieved.

A reflecting paraboloid must be carefully designed and constructed to obtain ideal performance. The antenna must be located at the focal point. The most desirable focal length of the parabola is that which places the radiator along the plane of the mouth: this length is equal to one-half the mouth radius. At other focal distances interference fields may deform the pattern or cancel a sizable portion of the radiation.

Corner Reflectors

The "corner" reflector consists of two flat conducting sheets which intersect at a designated angle. The corner-reflector antenna is particularly useful at v.h.f. where structures one or two wavelengths in maximum dimensions are more practical to build than larger systems.

The plane surfaces are set at an angle of 90 degrees, with the antenna set on a line bisecting this angle. For maximum performance, the distance of the antenna from the vertex should be 0.5 wavelength, but compromise designs can be built with closer spacings. The plane surfaces need not be solid sheets; spines spaced about 0.1 wavelength apart will serve as well. The spines do not have to be connected together electrically.

If the driven radiator is situated on a line bisecting the corner angle, as shown in Fig. 14-16, maximum radiation is in the direction of this line. There is no focus point for the driven radiator, as with a parabolic reflector, and the radiator can be placed at a variety of positions along the bisecting line.

Corner angles larger than 90 degrees can be used, with some decrease in gain. A 180-degree "corner" is equivalent to a single flat-sheet reflector. With angles smaller than 90 degrees, the gain theoretically increases as the corner angle is decreased. However, to realize this gain the size of the reflecting sheets must also be increased.

At a spacing of 0.5 wavelength from the driven dipole to the vertex, the radiation resistance of the driven dipole is approximately twice the radiation resistance of the same dipole in free space. Smaller spacings of driven dipole and vertex are practical, but at a slight sacrifice in efficiency. The alternative design for the 144- and 50-Mc. square-corner reflector has a dipole-to-vertex spacing of 0.4 wavelength. At this spacing the driven-dipole radiation resistance is still somewhat higher than its free-space value, but is considerably less than when the spacing is 0.5 wavelength.

U.H.F. and Microwaves

Once the amateur passes the 220-Mc. band on the way up through the radio-frequency spectrum, he encounters a distinct change of technique. So far he has been operating in a region where various modifications make usable the familiar coils and condensers, the crystal-controlled transmitters, selective superhet receivers, and other more-or-less standard items of the amateur field.

The boundary line beyond which such conventional gear is no longer usable has moved ever higher and higher in frequency as new developments and improvements in existing equipment have come along. In the early '30s the boundary line was our 28-Mc. band; then, as that band filled, the line moved up to 56 Mc., which remained border territory until 1938, when stabilization of transmitters used was made a legal requirement of operation in the old 5-meter band. For some years, then, the 112-Mc. band, and since the war the 144-Mc. band, constituted the dividing line, but even the latter band has now swung into the stabilized-transmitter-and-superhet-receiver field, and the 220-Mc. band is rapidly achieving the same status.

In the light of current developments, it may be said that the 420-Mc. band is now true borderline territory. The multistage transmitter can be used successfully, as can the superheterodyne receiver of semiconventional design, but special tank circuits must be employed and extreme care in mechanical layout must be used, in order to achieve satisfactory results.

The 420-Mc. band is fruitful territory for the experimentally-minded amateur. Most of the gear used will have to be made by the worker himself, but the techniques employed are such that construction of the necessary equipment will not be outside his capabilities. There is enough interest in a number of areas to support regular activity in this band, and more can be generated with a little organizational effort.

Antenna work on these frequencies is particularly intriguing. The antenna systems are so small in size that arrays having a gain of 10 db. or more can be erected in almost any location. Experimentation with models built for 420 Mc. is a fine way of checking the performance of arrays for lower frequencies. The experimenter who starts to work with u.h.f. antenna systems is bound to find himself spending many interesting hours checking his pet antenna ideas. Since u.h.f. or microwave experimentation is best accomplished in groups of interested workers, it is a fine project for coöperative effort by radio clubs.

The communication possibilities of the u.h.f. region should not be overlooked. Recent experience in the 144-Mc. band has demonstrated the possibilities of that band for long-distance work, and it is reasonable to assume that propagation vagaries, as regards tropospheric effects, will continue on up through the microwave range. With suitable antenna systems, it is probable that operating ranges on the frequencies above 200 Mc. may equal or approach those now being covered in the 70–160-Mc. region.

At least some amateur work has been done in all the microwave bands now assigned. The work of the pioneers in adapting these frequencies to communication purposes has been in line with the best amateur tradition, and it is hoped that the almost unknown territory from 500 Mc. up will see much amateur exploration in the near future.

U.H.F. Tank Circuits

In resonant circuits as employed at the lower frequencies it is possible to consider each of the reactance components as a separate entity. A coil is used to provide the required inductance and a condenser is connected across it to provide the needed capacitance. The fact that the coil itself has a certain amount of self-capacitance, as well as some resistance, while the condenser also possesses a small self-inductance, can usually be disregarded.

At the very-high and ultrahigh frequencies, however, it is no longer possible to separate these components. The connecting leads which, at lower frequencies, would serve merely to join the condenser to the coil now may have more inductance than the coil itself. The required inductance coil may be no more than a single turn of wire, yet even this single turn may have dimensions comparable to a wavelength at the operating frequency. Thus the energy in the field surrounding the "coil" may in part be radiated. At a sufficiently high frequency the loss by radiation may represent a major portion of the total energy in the circuit. Since energy which cannot be utilized as intended is wasted, regardless of whether it is

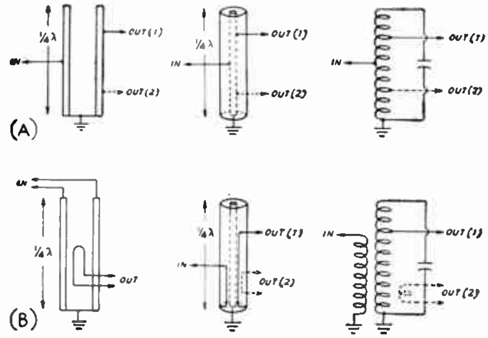


Fig. 15-1 — Equivalent coupling circuits for parallel-line, coaxial-line and conventional resonant circuits.

consumed as heat by the resistance of the wire or simply radiated into space, the effect is as though the resistance of the tuned circuit were greatly increased and its Q greatly reduced.

For this reason, it is common practice to utilize resonant sections of transmission line as tuned circuits at frequencies above 100 Mc. A quarter-wavelength line, or any odd multiple thereof, shorted at one end and open at the other, exhibits large standing waves. When a voltage of the frequency at which such a line is resonant is applied to the open end, the response is very similar to that of a parallel resonant circuit; it will have very high input impedance at resonance and a large current flowing at the short-circuited end. The input impedance may be as high as 0.4 megohm for a well-constructed line.

The action of a resonant quarter-wavelength line can be compared with that of a coil-and-condenser combination whose constants have been adjusted to resonance at a corresponding frequency. Around the point of resonance, in fact, the line will display very nearly the same characteristics as those of the tuned circuit. The equivalent relationships are shown in Fig. 15-1. At frequencies off resonance the line displays qualities comparable to the inductive and capacitive reactances of the coil-and-condenser circuit, although the exact relationships involved are somewhat different. For all practical purposes, however, sections of resonant wire or transmission line can be used in much the same manner as coils or condensers.

In circuits operating above 300 Mc., the spacing between conductors becomes an appreciable fraction of a wavelength. To keep the radiation loss as small as possible the parallel conductors should not be spaced farther apart than 10 per cent of the wavelength, center to center. On the other hand, the spacing of large-diameter conductors should not be reduced to much less twice the diameter because of what is known as the *proximity effect*, whereby another form of loss is introduced through eddy currents set up by the adjacent fields. Because the cancellation is no longer complete, radiation from an open line becomes so great that the Q is greatly reduced. Consequently, at these frequencies coaxial lines must be used.

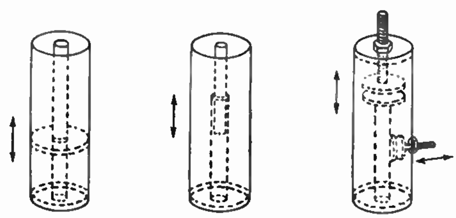


Fig. 15-2 — Methods of tuning coaxial resonant lines.

Construction

Practical information concerning the construction of transmission lines for such specific uses as feeding antennas and as resonant circuits in radio transmitters will be found in this and other chapters of this *Handbook*. Certain basic considerations applicable in general to resonant lines used as circuit elements may be considered here, however.

While either parallel-line or coaxial sections may be used, the latter are preferred for higher-frequency operation. Representative methods for adjusting the length of such lines to resonance are shown in Fig. 15-2. At the left, a sliding shorting disk is used to reduce the effective length of the line by altering the position of the short-circuit. In the center, the same effect is accomplished by using a telescoping tube in the end of the inner conductor to vary its length and thereby the effective length of the line. At the right, two possible methods of mounting parallel-plate condensers, used to tune a "foreshortened" line to resonance, are illustrated. The arrangement with the loading capacitor at the open end of the line has the greatest tuning effect per unit of capacitance; the alternative method, which is equivalent to "tapping" the condenser down on the line, has less effect on the Q of the circuit. Lines with capacitive "loading" of the sort illustrated will be shorter, physically, than an unloaded line resonant at the same frequency.

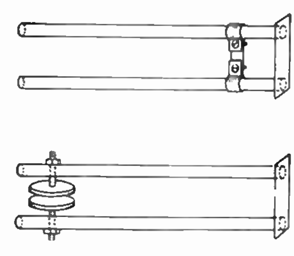


Fig. 15-3 — Methods of tuning parallel-type resonant lines.

The short-circuiting disk at the end of the line must be designed to make perfect electrical contact. The voltage is a minimum at this end of the line; therefore, it will not break down some of the thinnest insulating films. Usually a soldered connection or a tight clamp is used to secure good contact. When the length of line

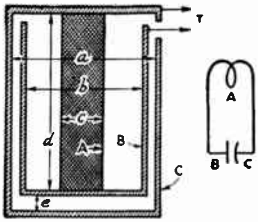


Fig. 15-4 — Concentric-cylinder or "pot"-type tank for v.h.f. The equivalent circuit diagram is also shown. Connections are made to the terminals marked *T*. For maximum *Q* the ratio of *b* to *c* should be between 3 and 5.

must be readily adjustable, the shorting plug is provided with spring collars which make contact on the inner and outer conductors at some distance away from the shorting plug at a point where the voltage is sufficient to break down the film between the collar and conductor.

Two methods of tuning parallel-conductor lines are shown in Fig. 15-3. The sliding short-circuiting strap can be tightened by means of screws and nuts to make good electrical contact. The parallel-plate condenser in the second drawing may be placed anywhere along the line, the tuning effect becoming less as the condenser is located nearer the shorted end of the line. Although a low-capacitance variable condenser of ordinary construction can be used, the circular-plate type shown is symmetrical and thus does not unbalance the line. It also has the further advantage that no insulating material is required.

Equivalent impedance points, for coupling or impedance-transformation purposes, are shown in Fig. 15-1 for parallel-line, coaxial-line, and conventional coil-and-condenser circuits.

Lumped-Constant Circuits

At the very-high frequencies the low values of L and C required make ordinary coils and condensers impracticable, while linear circuits offer mechanical difficulties in making tuning adjustments over a wide frequency range, and radiation from unshielded lines may reduce their effectiveness materially.

To overcome these difficulties, special high-Q lumped-constant circuits have been developed in which connections from the "condenser" to the "coil" are an inherent part of the structure. Integral design minimizes both resistance and inductance and increases the C/L ratio.

The simplest of these circuits is based on the use of disks combining half-turn inductance loops with semicircular condenser plates. By connecting several of these half-turn coils in parallel, the effective inductance is reduced to a value appreciably below that for a single turn. Tuning is accomplished by interleaving grounded rotor plates between the turns. Both by shielding action and short-circuited-turn effect, these further reduce the inductance.

Another type of high-C circuit is a single-turn toroid, commonly termed the "hat" resonator. Two copper shells with wide, flat "brims" are mounted facing each other on an axially-aligned copper rod. The capacitance in the circuit is that between the wide shells, while the central rod comprises the inductance.

"Pot"-Type Tank Circuits

The lumped-constant concentric-element tank in Fig. 15-4, commonly referred to as the "pot" circuit, is equivalent to a very short coaxial line (no linear dimension should exceed 1/20 wavelength), loaded by a large integral capacitor.

The inductance is supplied by the copper rod, *A*. Capacitance is provided by the concentric cylinders, *B* and *C*, plus the capacitance between the plates at the bottoms of the cylinders.

Approximate values of capacitance and inductance for tank circuits of the "pot" type can be determined by the following:

$$L = 0.0117\, d \log \frac{b}{c}\ \mu\text{h.}$$

$$C = \left(\frac{0.6128\, d}{\log \frac{a}{b}}\right) + \left(\frac{0.1775\, b^2}{e}\right)\ \mu\mu\text{fd.}$$

where the symbols are as indicated in Fig. 15-4, and dimensions are in inches. The left-hand term for capacitance applies to the concentric cylinders, *B* and *C*, while the second term gives capacitance between the bottom plates.

"Butterfly" Circuits

The tank circuits described in the preceding section are primarily fixed-frequency devices. The "butterfly" circuits shown in Fig. 15-5 are capable of being tuned over an exceptionally wide range, while still having high Q and reasonable physical dimensions. The circuit at A is derived from a conventional balanced-type variable condenser. The inductance is in the wide circular band connecting the stator plates. At its minimum setting the rotor plate fills the opening of the loop, reducing the inductance to a minimum. Connections are made to points *1* and *2*. This basic structure eliminates all connecting leads and avoids all sliding or wiping electrical contacts to a rotating member. A disadvantage is that the electrical midpoint shifts

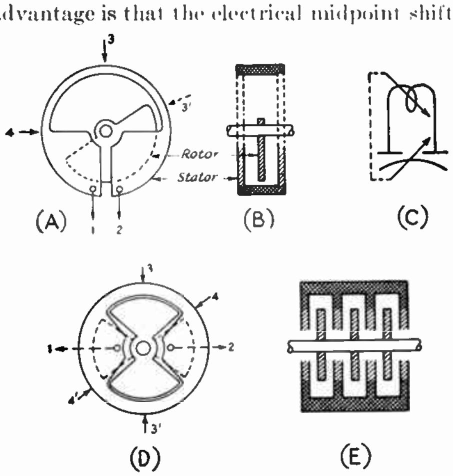


Fig. 15-5 — "Butterfly" tank circuits for v.h.f., showing front and cross-section views and the equivalent circuit.

from point *3* to point *3'* as the rotor is turned. Constant magnetic coupling may be obtained by a coupling loop located at point *4*, however.

In the modification shown at D, two sectoral stators are spaced 180 degrees, thereby achieving the electrical symmetry required to permit tapping for balanced operation. Connections to the circuit should be made at points *1* and *2* and it may be tapped at points *3* and *3'*, which are the electrical midpoints. Where magnetic coupling is employed, points *4* and *4'* are suitable locations for coupling links.

The capacitance of any butterfly circuit may be computed by the standard formula for parallel-plate condensers given in Chapter Twenty-Four. The maximum inductance can be obtained approximately by finding the inductance of a full ring of the same diameter and multiplying the result by a factor of 0.17. The ratio of minimum to maximum inductance varies between 1.5 and 4 with conventional construction.

Any number of butterfly sections may be connected in parallel. In practice, units of four to eight plates prove most satisfactory. The ring and stator sections may either be made in a single piece or with separate sectoral stator plates and spacing rings assembled with machine screws.

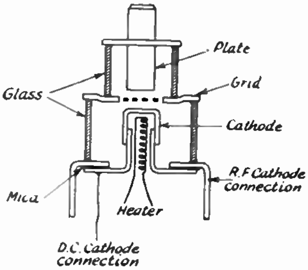


Fig. 15-6—**Sectional view of the "lighthouse" tube's construction. Close electrode spacing reduces transit time while the disk electrode connections reduce lead inductance.**

V.H.F. and U.H.F. Tubes

At very-high frequencies, interelectrode capacitance and the inductance of internal leads determine the highest possible frequency to which a vacuum tube can be tuned. The tube usually will not oscillate up to this limit, however, because of dielectric losses, grid emission, and "transit-time" effects. In low-frequency operation, the actual time of flight of electrons between the cathode and the anode is negligible in relation to the duration of the cycle. At 1000 kc., for example, transit time of 0.001 microsecond, which is typical of conventional tubes, is only 1/1000 cycle. But at 100 Mc., this same transit time represents 1/10 of a cycle and a full cycle at 1000 Mc. These limiting factors establish about 3000 Mc. as the upper frequency limit for negative-grid tubes.

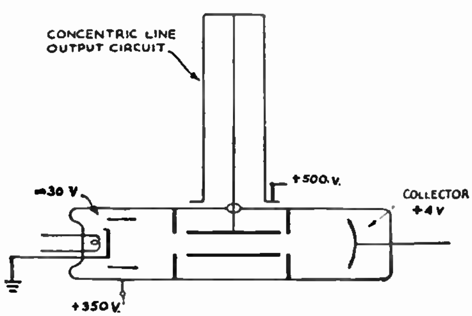


Fig. 15-7—**Simple form of cylindrical-grid velocity-modulated tube with retarding-field collector and coaxial-line output circuit, used as a superheterodyne high-frequency oscillator or as a superregenerative detector. Similar tubes can also be used as r.f. amplifiers and frequency converters in the 5–50-cm. region.**

With tubes of ordinary construction, the upper limit of oscillation is about 150 Mc. For higher frequencies, v.h.f. tubes of special construction are used. The "acorn" and "doorknob" types and the special v.h.f. "miniature" tubes, in which the grid-cathode spacing is made as little as 0.005 inch, are capable of operation up to about 700–800 Mc. The normal frequency limit is around 600 Mc., although output may be obtained up to 800 Mc.

Very low interelectrode capacitance and lead inductance have been achieved in the newer tubes of modified construction. In multiple-lead types the electrodes are provided with up to three separate leads which, when connected in parallel, have considerably-reduced effective inductance. In double-lead types the plate and grid elements are supported by heavy single wires which run entirely through the envelope, providing terminals at either end of the bulb. When a resonant circuit is connected to each pair of leads, the shunting capacitance divides between the two circuits. With linear circuits the leads become a part of the line and have distributed rather than lumped constants. Radiation loss is minimized and the effect of the transit time is reduced. In "lighthouse" tubes or *megatrons* the plate, grid and cathode are assembled in parallel planes, as shown in Fig. 15-6, instead of coaxially. The uniform coplanar electrode design and disk-seal terminals permit low interelectrode capacitance.

Velocity Modulation

In negative-grid operation the potential on the grid tends to reduce the electron velocity during the more negative half of the oscillation cycle, while on the other half-cycle the positive potential on the grid serves to accelerate them. Thus the electrons tend to separate into groups, those leaving the cathode during the negative half-cycle being collectively slowed down, while those leaving on the positive half are accelerated. After passing into the grid-plate space only a part of the electron stream follows the original form of the oscillation cycle, the remainder traveling to the plate at differing velocities. Since these contribute nothing to the

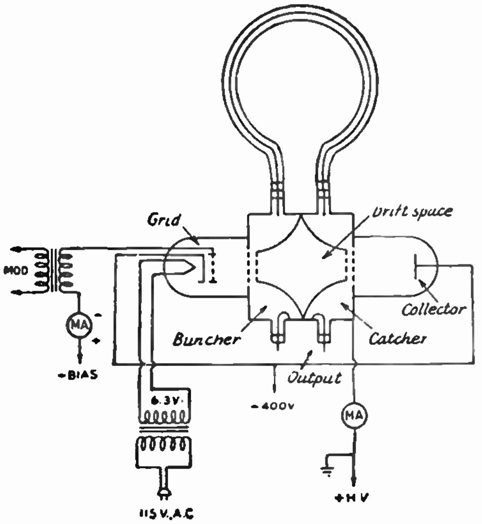


Fig. 15-8 — **Circuit diagram of the klystron oscillator, showing the feed-back loop coupling the frequency-controlling rhumbatrons and the output loop in the catcher.**

power output at the operating frequency, the efficiency is reduced in direct proportion to the variation in velocity, the output reaching a value of zero when the transit time approaches a half-cycle.

This effect, such a disadvantage in conventional tubes, is an advantage in velocity-modulated tubes in that the input signal voltage on the grid is used to change the velocity of the electrons in a constant-current electron beam, rather than to vary the intensity of a constant-velocity current flow as is the method in ordinary tubes.

A simple form of velocity-modulation oscillator tube is shown in Fig. 15-7. Electrons emitted from the cathode are accelerated through a negatively-biased cylindrical grid by a constant positive voltage applied to a sleeve electrode, shown in heavy lines. This electrode, which is the velocity-modulation control grid, consists of two hollow tubes, with a small space at each end between the inner tube, through which the electron beam passes, and the disks at the ends of the larger tube portion. With r.f. voltage applied across these gaps, which are small compared to the distance traveled by the electrons in one half-cycle, electrons entering the tube will be accelerated on positive half-cycles and decelerated on the negative half-cycles. The length of the tube is made equal to the distance covered by the electrons in one-half cycle, so that the electrons will be further accelerated or decelerated as they leave the tube.

As the beam approaches the collector electrode, which is at nearly zero potential, the electrons are retarded, brought to rest, and ultimately turned back by the attraction of the positive sleeve electrode. The collector electrode is, therefore, also termed a *reflector*. The point at which electrons are returned depends on their velocity. Thus the velocity modulation is again translated into current modulation.

Velocity-modulated tubes operate satisfactorily up to 6000 Mc. (5 cm.) and higher, with outputs of 100 watts or more.

The Klystron

In the *klystron* velocity-modulated tube, the electrons emitted by the cathode are accelerated or retarded during their passage through an electric field established by two grids in a cavity resonator, or *rhumbatron,* called the "buncher." The high-frequency electric field between the grids is parallel to the electron stream. This field accelerates the electrons at one moment and retards them at another, in accordance with the variations of the r.f. voltage applied. The resulting velocity-modulated beam travels through a field-free "drift space," where the slowly-moving electrons are gradually overtaken by the faster ones. The electrons emerging from the pair of grids therefore are separated into groups or bunched along the direction of motion. The velocity-modulated electron stream is passed to a "catcher" rhumbatron. Again the beam passes through two parallel grids; the r.f. current created by the bunching of the electron beam induces an r.f. voltage between the grids. The catcher cavity is made resonant at the frequency of the velocity-modulated electron beam, so that an oscillating field is set up within it by the passage of the electron bunches through the grid aperture.

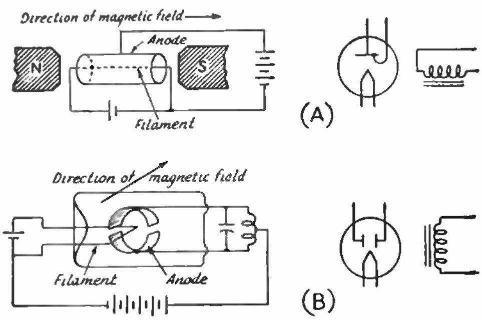


Fig. 15-9 — **Conventional magnetrons, with equivalent schematic symbols at the right. A, simple cylindrical magnetron. B, split-anode negative-resistance magnetron.**

If a feed-back loop is provided between the two rhumbatrons, as shown in Fig. 15-8, oscillations will occur. The resonant frequency depends on the electrode voltages and on the shape of the cavities, and may be adjusted by varying the supply voltage and altering the dimensions of the rhumbatrons. The bunched beam current is rich in harmonics, but the output waveform is remarkably pure because the high Q of the catcher rhumbatron suppresses the unwanted harmonics.

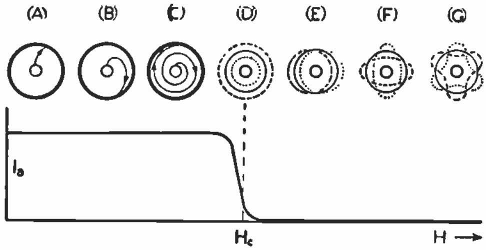


Fig. 15-10 — Electron trajectories for increasing values of magnetic field strength, *H*. Below is shown the corresponding curve of plate current, I_a. Oscillations commence when *H* reaches a critical value, H_c; progressively higher-order modes of oscillation occur beyond this point.

Magnetrons

A magnetron is fundamentally a diode with cylindrical electrodes placed in a uniform magnetic field with the lines of electromagnetic force parallel to the elements. The simple cylindrical magnetron consists of a filamentary cathode surrounded by a concentric cylindrical anode. In the more efficient split-anode magnetron the cylinder is divided longitudinally.

Magnetron oscillators are operated in two different ways. Electrically the circuits are similar, the difference being in the relation between electron transit time and the frequency of oscillation.

In the negative-resistance or dynatron type of magnetron oscillator, the element dimensions and anode voltage are such that the transit time is short compared with the period of the oscillation frequency. Electrons emitted from the cathode are driven toward both halves of the anode. If the potentials of the two halves are unequal, the effect of the magnetic field is such that the majority of the electrons travel to that half of the anode that is at the lower potential. In other words, a *decrease* in the potential of either half of the anode results in an *increase* in the electron current flowing to that half. The magnetron consequently exhibits negative-resistance characteristics. Negative-resistance magnetron oscillators are useful between 100 and 1000 Mc. Under the best operating conditions efficiencies of 20 to 25 per cent may be obtained. Since the power loss in the tube appears as heat in the anode, where it is readily dissipated, relatively large power-handling capacity can be obtained.

In the transit-time magnetron the frequency is determined primarily by its dimensions and

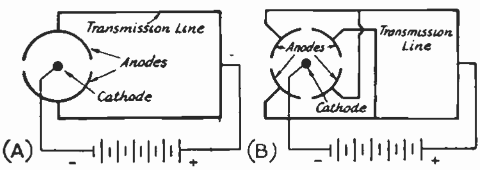


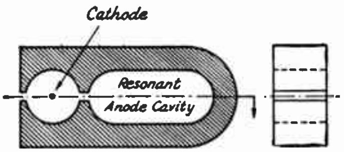
Fig. 15-11 — S.h.f. magnetron circuits. A, split-anode type. B, 4-anode type, opposite electrodes paralleled.

by the electric and magnetic field intensities rather than by the tuning of the tank circuits. The efficiency is much better than that of a positive-grid oscillator and good power output can be obtained even on the superhighs.

In a nonoscillating magnetron with a weak magnetic field, electrons traveling from the cathode to the anode move almost radially, their trajectories being bent only slightly by the magnetic field. With increased magnetic field the electrons tend to spiral around the filament, their radial component of velocity being much smaller than the angular component. Under critical conditions of magnetic field strength, a cloud of electrons rotates about the filament. It extends up to the anode but does not actually reach it.

The nature of these electron trajectories is shown in Fig. 15-10. Cases A, B and C correspond to the nonoscillating condition. For a small magnetic field (A) the trajectory is bent slightly near the anode. This bending increases for a higher magnetic field (B) and the electron moves through quite a large angle near the anode before reaching it, signifying a large increase of space charge near the anode. For a

Fig. 15-12 — Split-anode magnetron with integral resonant anode cavity for use at u.h.f.



strong magnetic field (C) electrons start radially from the cathode but are soon bent and curl about the filament in the form of a long spiral before reaching the anode. This means a very long transit time and a very large space charge in the whole region where the spiraling takes place. Under critical conditions (D), no current flows to the anode and no electron is able to move from cathode to anode, but a large space charge still exists between the cathode and anode. The spiraling becomes a set of concentric circles, and the entire space-charge distribution rotates about the filament.

Fig. 15-10E, F and G depicts higher-order (harmonic-type) modes of operation in which the space charge oscillates not only symmetrically but in transverse directions contrasting to the vibrations of the fundamental.

In a transit-time magnetron oscillator the intensity of the magnetic field is adjusted so that, under static conditions, electrons leaving the cathode move in curved paths which just fail to reach the anode. All electrons are therefore deflected back to the cathode, and the anode current is zero. When an alternating voltage is applied between the two halves of the anode, causing the potentials of these halves to vary about their average positive values, the conditions in the tube become analogous to those in a positive-grid oscillator. If the period of the alternating voltage is made equal to the time required for an electron to make one complete rotation in the magnetic field, the a.c. component of the anode voltage reverses direction twice with each electron rotation. Some electrons will lose energy to the electric field, with the result that they are unable to reach the cathode and continue to rotate about it. Meanwhile other electrons gain energy from the field and are returned to the cathode. Since those electrons that lose energy remain in the interelectrode space longer than those that gain energy, the net effect is a transfer of energy from the electrons to the electric field. This energy can be applied to sustain oscillations in a resonant transmission line connected between the two halves of the anode.

Split-anode magnetrons for u.h.f. are constructed with a cavity resonator built into the tube structure, as illustrated in Fig. 15-12. The assembly is a solid block of copper which assists in heat dissipation. At extremely high frequencies operation is improved by subdividing the anode structure into from 4 to 16 or more segments, the resonant cavities for each anode coupled by slots of critical dimensions to the common cathode region, as in Fig. 15-13.

The efficiency of multisegment magnetrons reaches 65 or 70 per cent. Slotted-anode magnetrons with four segments function up to 30,000 Mc. (1 cm.), delivering up to 100 watts at efficiencies greater than 50 per cent. Using larger multiples of anodes and higher-order modes, performance can be attained at 0.2 cm.

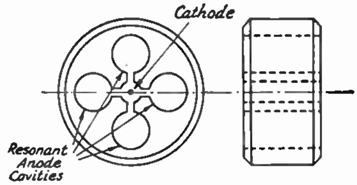


Fig. 15-13 — Multisegment magnetron with four resonant cavities. This construction is used for extremely high frequencies.

Equipment for 420 Mc.

Though it is possible to use crystal control on 420 Mc. it is improbable that all amateurs will care to go to the trouble necessary to accomplish it. The same is true in reception of 420-Mc. signals; many workers will be looking for the simpler forms of gear, in the early phases of their 420-Mc. endeavor. Thus we may expect that, for some time to come, much of the work on this band will be done with simple oscillator-type transmitters and superregenerative receivers.

The next step up from the superregenerator is the superheterodyne converter or receiver using a very broad i.f. amplifier, such as is employed in radar service. This sort of i.f. system is readily adapted to amateur use, and its extreme broadness is not a serious handicap at the present state of activity on this frequency. Lack of selectivity in the receiver, and broadness of the signal radiated by simple transmitters, while not conducive to best results, need not be troublesome from the standpoint of interference, as the 420-Mc. band is much wider than any lower amateur assignment.

SIMPLE 420-MC. GEAR FOR THE BEGINNER

The 420-Mc. transmitter and receiver shown in Figs. 15-14 to 15-21 are about the simplest equipment with which satisfactory communication can be carried on. Both employ 6J6 tubes in their r.f. portions, the circuits being

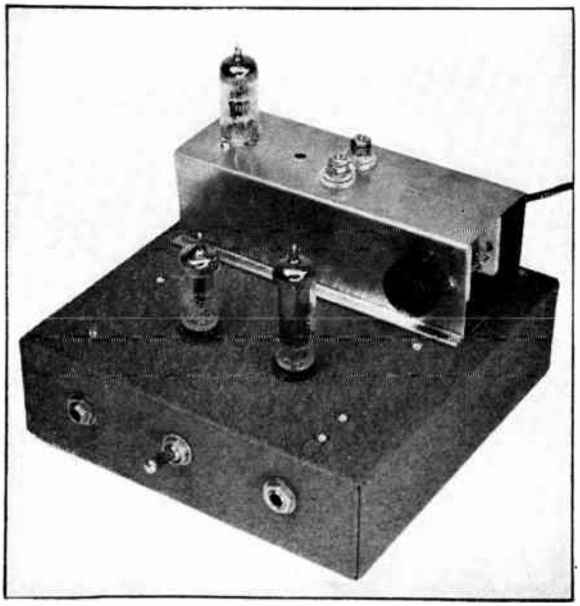


Fig. 15-14 — A 420-Mc. transmitter built in two units. The modulator portion, on a 7 × 7 × 2-inch chassis, uses a 6C4 driving a 6AQ5 modulator. The oscillator uses a 6J6 and is assembled on a removable trough-shaped chassis.

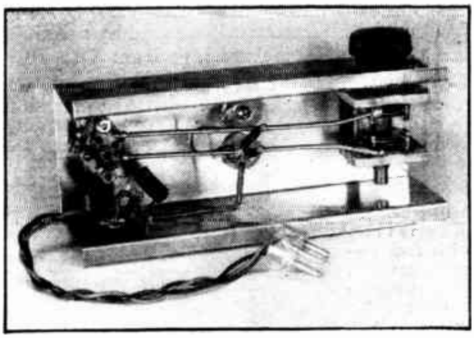


Fig. 15-15 — Bottom view of the oscillator assembly. The trough in which the components are mounted is made of flashing copper. It is 6 inches long, $1\frac{7}{8}$ inches high, and $2\frac{1}{4}$ inches wide, with $\frac{1}{4}$-inch edges folded over for sliding into a clip attached to the main chassis.

practically identical schematically. The tuned circuit in each is a half-wave line, with the tube plates at one end and the tuning condenser at the other. The plate voltage is fed into the line at the approximate middle, the exact point being determined by experiment. Two 100-ohm resistors, R_7 and R_8 in Fig. 15-16, are used at the feed point in the transmitter, as a precaution against loss of r.f. into the power-supply lead. The receiver uses a small center-tapped choke, RFC_1 in Fig. 15-20, for this purpose, and a similar arrangement may be used in the transmitter, if desired. The only other oscillator circuit difference between the two units is the value of the grid leak, and the use in the receiver of the by-pass condenser C_1 in the grid lead to induce superregeneration. The cathode and heater are maintained above ground potential in both units by small self-supporting r.f. chokes.

The audio portions of the receiver and transmitter are also quite similar circuitwise. In the transmitter a 6C4 speech amplifier is operated with the microphone-transformer primary connected in its cathode lead, thus doing away with the necessity for a microphone battery. This drives a 6AQ5, providing more than enough output for modulating the 5 or 6 watts input to the 6J6 oscillator. The receiver audio system uses a 6J5 and a 6F6.

Mechanical Details

The secret of success in getting the 6J6 tubes to operate satisfactorily at 420 Mc. lies in the elimination of all "leads" in the radio-frequency circuits. The plate line, L_2, is connected directly to the socket pins, as are the grid resistors and the heater chokes. Use of the half-wave line, in place of the more common capacitance-loaded quarter-wave arrangement, permits the use of a standard readily-obtainable tuning condenser, yet leaves a line of appreciable length. Using half-wave lines in the manner shown the 6J6 can be made to oscillate up to 700 Mc. or more with ease.

The oscillator portion of the transmitter, Fig. 15-15, is built inside a trough made of flashing copper, which is easy to work with simple tools and ideal from the standpoint of conductivity and shielding qualities. It is inexpensive and may be obtained from building-supply houses everywhere. The trough is fitted to a copper clip fastened to the main chassis. Power connections are made with a small plug and socket, the latter being mounted on the rear wall of the main chassis. This permits experimentation with the oscillator portion, or even substitution of r.f. sections for other bands, without the necessity for making changes in the modulator unit. This trough

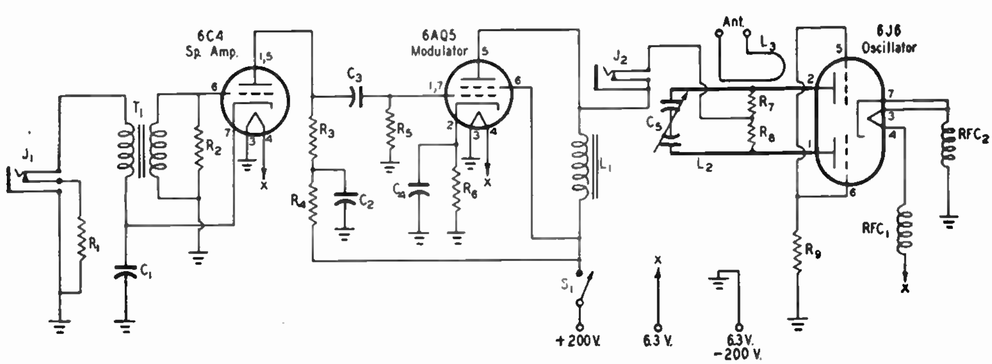


Fig. 15-16 — Schematic diagram of the 420-Mc. transmitter.

C_1, C_4 — 10-μfd. 25-volt electrolytic.
C_2 — 8-μfd. 450-volt electrolytic.
C_3 — 0.01-μfd. tubular.
C_5 — Miniature split-stator variable, 4 $\mu\mu$fd. per section. (Millen 21912D, with one rotor plate removed from each section.)
R_1 — 470 ohms, 1 watt.
R_2 — 0.33 megohm, $\frac{1}{2}$ watt.
R_3, R_4 — 5000 ohms, 5 watts.
R_5 — 0.47 megohm, $\frac{1}{2}$ watt.
R_6 — 680 ohms, 1 watt.
R_7, R_8 — 100 ohms, $\frac{1}{2}$ watt, carbon.
R_9 — 2700 ohms, $\frac{1}{2}$ watt.
L_1 — Midget filter choke.
L_2 — Plate line made of two pieces of No. 12 wire, $4\frac{1}{4}$ inches long, $\frac{3}{8}$ inch apart, center to center.
L_3 — Hairpin of No. 18 wire. Portion which couples to L_2 is about $\frac{5}{8}$ inch long. Position should be adjusted for maximum transfer of power to antenna.
J_1, J_2 — Closed-circuit jack.
RFC_1, RFC_2 — 12 turns No. 20 enameled wire, $\frac{3}{16}$-inch diam., $\frac{3}{4}$ inch long.
T_1 — Single-button microphone transformer.

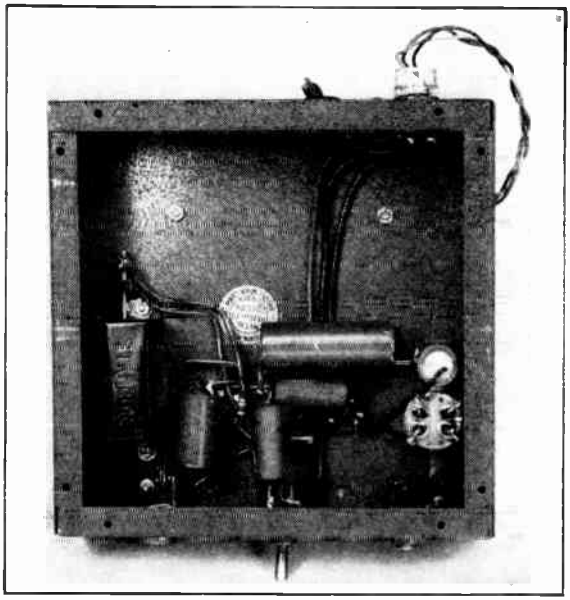


Fig. 15-17 — Bottom view of the main chassis of the 420-Mc. transmitter, showing audio components.

construction also helps prevent direct radiation from the tank circuit. The useful output with this type of assembly is nearly twice that obtainable with open construction.

The 6J6 plate line in the receiver, Fig. 15-19, is bent in the shape of an inverted "U," with the tube socket mounted on a small bracket near the edge of the chassis. A padder adjustment is added in the form of two copper plates soldered to the stator terminals of C_6. These are approximately ½ by 1 inch in size, and are bent toward one another until the desired setting of the band is obtained.

The antenna coupling loop should be shaped so that it may be placed parallel to the plane of the line at a position about ⅛ to ¼ of an inch above it. As the frequency range to be covered is considerable, the degree of loading by the antenna varies widely over the band, and some form of adjustable antenna coupling is an absolute necessity. Don't try to do without it — the detector cannot be made to operate at maximum sensitivity unless the coupling is adjusted with extreme care.

Testing

Lecher Wires provide the best means of checking the frequency of the 420-Mc. transmitter. Information on the construction of Lecher Wires may be found in Chapter Sixteen. After the transmitter range has been calibrated by this means, an absorption-type wavemeter may be made and used thereafter for approximate checks. The tuning range of the superregenerative receiver may be checked with either device, or it may be done by listening to the transmitter, once the frequency of that unit has been determined.

A convenient absorption-type wavemeter may be made by bending 6 inches of Number

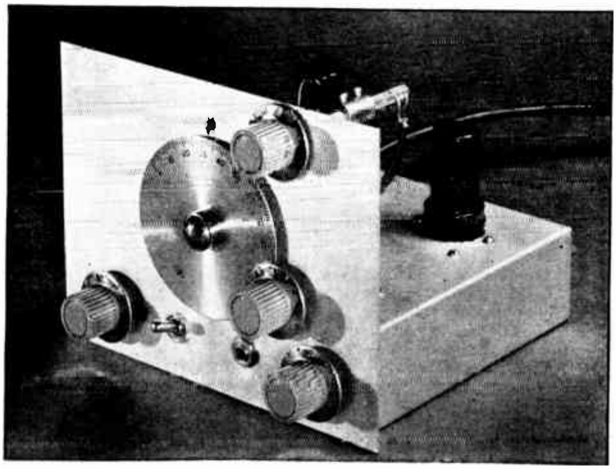


Fig. 15-18 — A superregenerative receiver for 420 Mc. The two lower controls are for variation of detector voltage (left) and audio gain. The vernier dial is the main tuning and the knob at the top adjusts the antenna coupling.

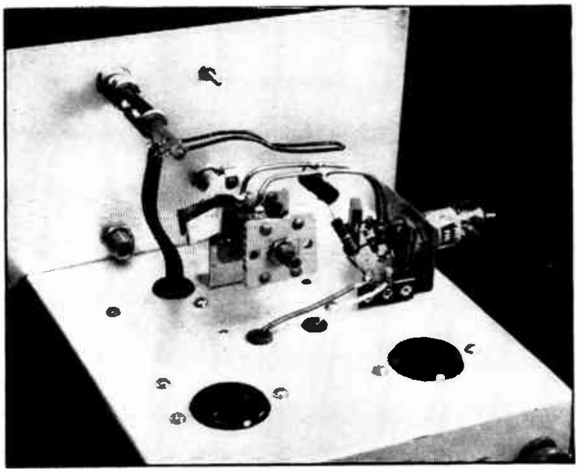


Fig. 15-19 — Detail view of the 120-Mc. superregenerative receiver. Note the method of varying the antenna coupling. Copper plates attached to the tuning-condenser stators provide a bandset adjustment.

12 wire into a "U" 1⅛ inches across, and soldering its ends to the stator terminals of a 2-plate Cardwell Trim-Aire, the stator plate of which has been sawed down the middle. The rotor plate should be spaced 1/16 inch from the split-stator plate. This wavemeter will spread the 420-Mc. band over three-quarters of its tuning range.

The grid current of the transmitter provides a good check on its performance. This may be measured by inserting a meter between R_9 and ground. With 200 volts on the plate the grid current should run about 5 to 6 ma. under load. A suitable load for the transmitter is a 6-8-volt 150-ma. pilot lamp, which should show a full-brilliance indication at about 30 ma. plate current.

Adjustment of the antenna coupling will probably be different with the antenna than with a lamp load, so this adjustment should be made with the antenna connected. A simple field-strength indicator may be a folded dipole 12¾ inches long, with a 60-ma. pilot lamp, or a 1N34 crystal and a microammeter, connected at its center. The antenna coupling loop may be adjusted by means of a fiber crochet hook inserted through a small hole in the top of the trough.

For most efficient operation, the point of connection of the two resistors, R_7 and R_8,

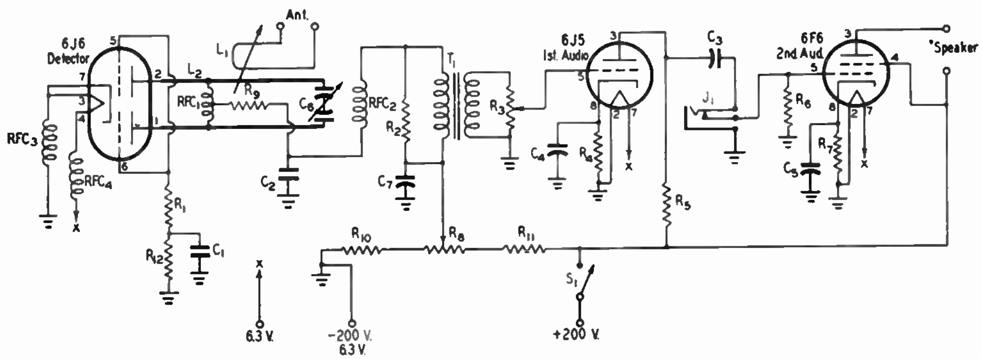


Fig. 15-20 — Schematic diagram of the 120-Mc. superregenerative receiver.

C_1 — 470-μμfd. mica.
C_2 — 0.0033-μfd. mica.
C_3 — 0.01-μfd. tubular.
C_4, C_5 — 10-μfd. 25-volt electrolytic.
C_6 — Miniature split-stator variable, about 4 μμfd. per section. (Millen 21912D, with one rotor plate removed from each section. See text and photo.)
C_7 — 0.1-μfd. tubular.
R_1 — 3800 ohms, ½ watt.
R_2 — 47,000 ohms, ½ watt.
R_3 — 0.5-megohm potentiometer.
R_4 — 2200 ohms, 1 watt.
R_5, R_6 — 0.1 megohm, ½ watt.
R_7 — 470 ohms, 1 watt.
R_8 — 50,000-ohm potentiometer.
R_9 — 2200 ohms, 1 watt.
R_{10}, R_{11} — 47,000 ohms, 1 watt.
R_{12} — 1.5 megohms, ½ watt.
L_1 — Hairpin loop No. 14 enameled wire, same spacing as L_2. Connect to antenna terminals by means of 300-ohm line.
L_2 — Half-wave line No. 12 wire, each side 3½ inches long, spaced ⅜ inch center to center. (See text and photographs for other details of L_1 and L_2.)
J_1 — Closed-circuit jack.
RFC_1 — 19 turns No. 20 enameled wire, 3/16-inch inside diameter, ⅞ inch long, center-tapped.
RFC_2 — 10-mh. r.f. choke.
RFC_3, RFC_4 — 12 turns No. 20 enameled wire, 3/16-inch inside diameter, ¾ inch long.
S_1 — S.p.s.t. toggle switch.
T_1 — Interstage audio transformer.

should be adjusted carefully. Starting with the connection near the middle of the line, touch a pencil along the line in either direction, noting the grid current meanwhile. A spot will be found where there is little or no change in grid current when this is done. This is the spot for the B-plus connections. If an appreciable change in the point of connection is made the frequency of the oscillator should be checked again.

The receiver should be checked in a similar manner, except that listening to the receiver will replace observation of the grid current as the pencil test is made. The position of the antenna coupling loop in the receiver will be found to be quite critical, and the best position will change in tuning across the band. The optimum setting will be that just before the detector goes out of oscillation as the antenna coupling is increased. Both the antenna-coupling and the regeneration-control settings will affect the frequency of the receiver, so it will be necessary to retune the receiver as these are adjusted. The copper plates attached to the stator plates of the tuning condenser provide a means of adjusting the bandspread and the position of the band on the dial.

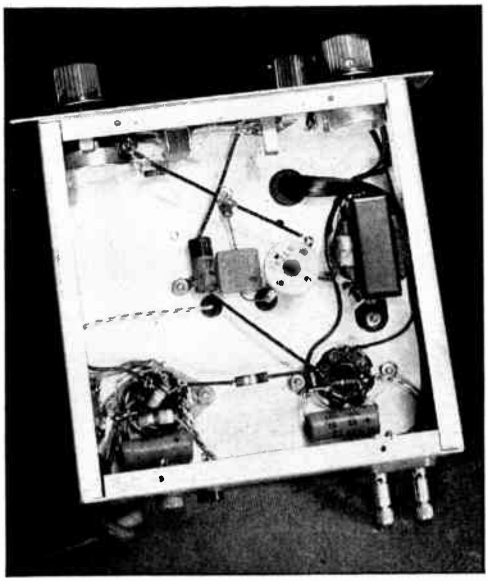


Fig. 15-21 — Bottom view of the 420-Mc. receiver. Loudspeaker terminals are at the lower left. At the right are the antenna terminals, from which a length of 300-ohm line runs up through the chassis to the antenna coupling loop.

Bibliography on 420-Mc. Equipment

For the convenience of the experimenter who is interested in 420 Mc. a list of articles appearing in *QST* and the *Handbook* since 1946 is given below.

"Getting Started on 420 Mc." (Hoisington), June 1946 *QST*, page 43.

"Four-Twenty Is Fun" (Tilton), Nov. 1947 *QST*, page 13.

"Operating the BC-645 on 420 Mc." (Ralph and Wood), Feb. 1947 *QST*, page 15.

"Fun on 420 with the BC-788" (Clapp), July 1948 *QST*, page 21.

"Operating the APS-13 on 420 Mc." (Addison), May 1948 *QST*, page 57.

"Tripling to 420 Mc." (Brannin), June 1948 *QST*, page 52.

"A Doorknob Oscillator for 420 Mc." (Tilton), January 1949 *QST*, page 29.

"Simple Gear for the 420-Mc. Beginner" (Tilton), May 1949 *QST*, page 11.

"An Acorn Converter for 420 Mc.," 1949 ARRL *Handbook*, page 452.

Wave Guides and Cavity Resonators

A wave guide is a conducting tube through which energy is transmitted in the form of electromagnetic waves. The tube is not considered as carrying a current in the same sense that the wires of a two-conductor line do, but rather as a *boundary* which confines the waves to the enclosed space. Skin effect prevents any electromagnetic effects from being evident outside the guide. The energy is injected at one end, either through capacitive or inductive coupling or by radiation, and is received at the other end. The wave guide then merely confines the energy of the fields, which are propagated through it to the receiving end by means of reflections against its inner walls.

The difficulty of visualizing energy transfer without the usual closed circuit can be relieved somewhat by considering the guide as being evolved from an ordinary two-conductor line.

In Fig. 15-22A, several closed quarter-wave stubs are shown connected in parallel across a two-wire transmission line. Since the open end of each stub is equivalent to an open circuit, the line impedance is not affected by their presence. Enough stubs may be added to form a "U"-shaped rectangular tube with solid walls, as at B, and another identical "U"-shaped tube may be added edge-to-edge to form the rectangular

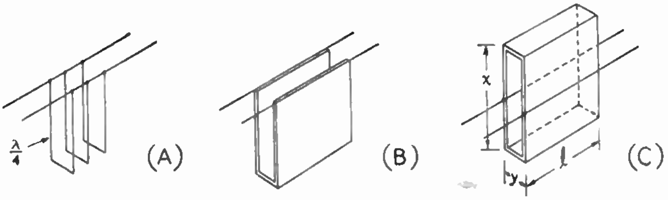


Fig. 15-22 — Evolution of a wave guide from a two-wire transmission line.

pipe shown in Fig. 15-22C. As before, the line impedance still will not be affected. But now, instead of a two-wire transmission line, the energy is being conducted within a hollow rectangular tube.

This analogy to wave-guide operation is not exact, and therefore should not be taken too literally. In the evolution from the two-wire line to the closed tube the electric- and magnetic-field configurations undergo considerable change, with the result that the guide does not actually operate like a two-conductor line shunted by an infinite number of quarter-wave stubs. If it did, only waves of the proper length to correspond to the stubs would be propagated through the tube, but the fact is that such waves do *not* pass through the guide. Only waves of shorter length — that is, higher frequency — can go through. The distance x represents half the *cut-off wavelength*, or the shortest wavelength that is unable to go through the guide. Or, to put it another way, waves of length equal to or greater than $2x$ cannot be propagated in the guide.

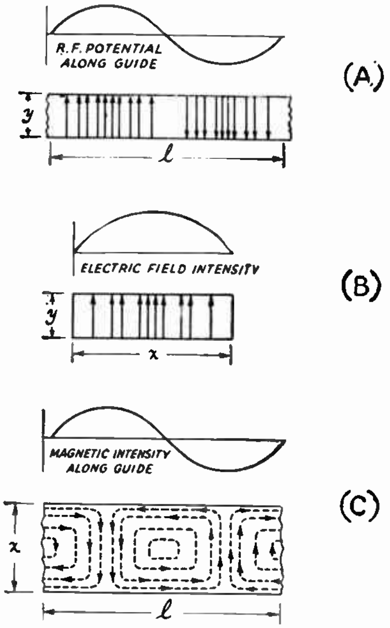


Fig. 15-23 — Field distribution in a rectangular wave guide. The $TE_{1,0}$ mode of propagation is depicted.

A second point of difference is that the apparent length of a wave along the direction of propagation through a guide always is greater than that of a wave of the same frequency in free space, whereas the wavelength along a two-conductor transmission line is the same as the free-space wavelength (when the insulation between the wires is air).

Operating Principles of Wave Guides

Analysis of wave-guide operation is based on the assumption that the guide material is a perfect conductor of electricity. Typical distributions of electric and magnetic fields in a rectangular guide are shown in Fig. 15-23. It will be observed that the intensity of the electric field is greatest at the center along the x dimension, diminishing to zero at the end walls. The latter is a necessary condition, since the existence of any electric field parallel to the walls at the surface would cause an infinite current to flow in a perfect conductor. This represents an impossible situation.

Zero electric field at the end walls will result if the wave is considered to consist of two separate waves moving in zigzag fashion down the guide, reflected back and forth from the end walls as shown in Fig. 15-24. Just at the walls, the positive crest of one wave meets the negative crest of the other, giving complete cancellation of the electric fields. The angle of reflection at which this cancellation occurs depends upon the width x of the guide and the length of the waves; Fig. 15-24A illustrates the case of a wave considerably shorter than the cut-off wavelength, while B shows a longer wave. When the wavelength equals the cut-off value, the two waves simply bounce back and forth between the walls and no energy is transmitted through the guide.

The two waves travel with the speed of light, but since they do not travel in a straight line the energy does not travel through the guide as rapidly as it does in space. A further consequence of the repeated reflections is that the points of maximum intensity or wave crests are separated more along the line of propagation in the guide than they are in the two separate waves. In other words, the wavelength in the guide is greater than the free-space wavelength. This is also shown in Fig. 15-24.

Modes of Propagation

Fig. 15-23 represents a relatively simple distribution of the electric and magnetic fields. There is in general an infinite number of ways in which the fields can arrange themselves in a guide so long as there is no upper limit to the frequency to be transmitted. Each field configuration is called a *mode*. All modes may be separated into two general groups. One group, designated *TM* (*transverse magnetic*), has the

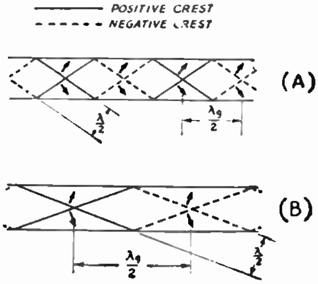


Fig. 15-24 — Reflection of two component waves in a rectangular guide. λ = wavelength in space, λg = wavelength in guide. Direction of wave motion is perpendicular to the wave front (crests) as shown by the arrows.

magnetic field entirely transverse to the direction of propagation, but has a component of electric field in that direction. The other type, designated *TE* (*transverse electric*) has the electric field entirely transverse, but has a component of magnetic field in the direction of propagation. *TM* waves are sometimes called *E* waves, and *TE* waves are sometimes called *H* waves, but the *TM* and *TE* designations are preferred.

The particular mode of transmission is identified by the group letters followed by two subscript numerals: for example, $TE_{1,0}$, $TM_{1,1}$, etc. The number of possible modes increases with frequency for a given size of guide. There is only one possible mode (called the *dominant* mode) for the lowest frequency that can be transmitted. The dominant mode is the one generally used in practical work.

Wave-Guide Dimensions

In the rectangular guide the critical dimension is x in Fig. 15-22; this dimension must be more than one-half wavelength at the lowest frequency to be transmitted. In practice, the y dimension usually is made about equal to $\frac{1}{2}x$ to avoid the possibility of operation at other than the dominant mode.

Other cross-sectional shapes than the rectangle can be used, the most important being the circular pipe. Much the same considerations apply as in the rectangular case.

Wavelength formulas for rectangular and circular guides are given in the following table, where x is the width of a rectangular guide and r is the radius of a circular guide. All figures are in terms of the dominant mode.

	Rectangular	*Circular*
Cut-off wavelength	$2x$	$3.41r$
Longest wavelength transmitted with little attenuation	$1.6x$	$3.2r$
Shortest wavelength before next mode becomes possible	$1.1x$	$2.8r$

Cavity Resonators

At low and medium radio frequencies resonant circuits usually are composed of "lumped" constants of L and C; that is, the inductance is concentrated in a coil and the capacitance concentrated in a condenser. However, as the frequency is increased, coils and condensers must be reduced to impracticably small physical dimensions. Up to a certain point this difficulty may be overcome by using linear circuits but even these fail at extremely high frequencies. Another kind of circuit particularly applicable at wavelengths of the order of centimeters is the *cavity resonator*, which may be looked upon as a section of a wave guide with the dimensions chosen so that waves of a given length can be maintained inside.

The derivation of one type of cavity resonator from an ordinary *LC* circuit is shown in Fig. 15-25. As in the case of the wave-guide derivation, this picture must be accepted with some reservations, and for the same reasons.

Considering that even a straight piece of wire has appreciable inductance at very-high frequencies, it may be seen in Fig. 15-25A and B that a direct short across a two-plate condenser with air dielectric is the equivalent of a tuned circuit with a typical coiled inductance. With two wires between the plates, as shown in Fig. 15-25C, the circuit may be thought of as a resonant-line section. For d.c. or even low frequency r.f., this line would appear as a short across the two condenser plates. At the ultra-high frequencies, however, such a section of line a quarter wavelength long would appear as an open circuit when viewed from one of the plates with respect to the other end of the section.

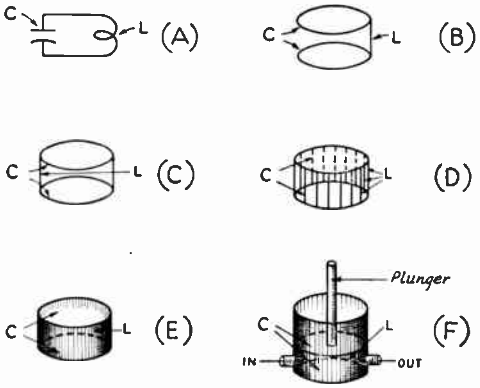


Fig. 15-25 — Steps in the derivation of a cavity resonator from a conventional coil-and-condenser tuned circuit.

Increasing the number of parallel wires between the plates of the condenser would have no effect on the equivalent circuit, as shown at D. Eventually, the closed figure at E will be developed. Since each wire which is added in D is like connecting inductances in parallel, the total inductance across the condenser becomes increasingly smaller as the solid form is approached, and the resonant frequency of the figure therefore becomes higher.

If energy now is introduced into the cavity in a manner such as that shown at F, the circuit will respond like any equivalent coil-condenser tank circuit at its resonant frequency. A cavity resonator may therefore be used as a u.h.f. tuning element, along with a vacuum tube of suitable design, to form the main components of an oscillator circuit which will be capable of functioning at frequencies considerably beyond the maximum limits possible when conventional tubes, coils and condensers are employed.

Other shapes than the cylinder may be used as resonators, among them the rectangular box, the sphere, and the sphere with re-entrant cones, as shown in Fig. 15-26. The resonant frequency depends upon the dimensions of the cavity and the mode of oscillation of the waves (comparable to the transmission modes in a wave guide). For the lowest modes the resonant wavelengths are as follows:

Cylinder	2.61r
Square box	1.41l
Sphere	2.28r
Sphere with re-entrant cones	4r

The resonant wavelengths of the cylinder and square box are independent of the height when the height is less than a half-wavelength. In other modes of oscillation the height must be a multiple of a half-wavelength as measured inside the cavity. Fig. 15-25F shows how a cylindrical cavity can be tuned when operating in such a mode. Other tuning methods include placing adjustable tuning paddles or "slugs" inside the cavity so that the standing-wave pattern of the electric and magnetic fields can be varied.

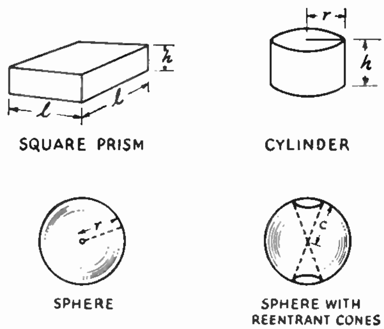


Fig. 15-26 — **Forms of cavity resonators.**

A form of cavity resonator in wide practical use is the re-entrant cylindrical type shown in Fig. 15-27. It is useful in connection with vacuum-tube oscillators of the types described for u.h.f. use earlier in this chapter. In construction it resembles a concentric line closed at both ends with capacitance loading at the top, but the actual mode of oscillation may differ considerably from that occurring in coaxial lines. The resonant frequency of such a cavity depends upon the diameters of the two cylinders and the distance d between the ends of the inner and outer cylinders.

Compared to ordinary resonant circuits, cavity resonators have extremely-high Q. A value of Q of the order of 1000 or more is readily obtainable, and Q values of several thousand can readily be secured with good design and construction.

Coupling to Wave Guides and Cavity Resonators

Energy may be introduced into or abstracted from a wave guide or resonator by means of either the electric or magnetic field. The energy transfer frequently is through a coaxial line, two methods for coupling to which

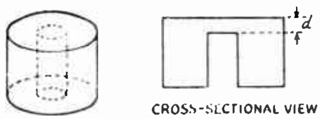


Fig. 15-27 — **Re-entrant cylindrical cavity resonator.**

are shown in Fig. 15-28. The probe shown at A is simply a short extension of the inner conductor of the coaxial line, so oriented that it is parallel to the electric lines of force. The loop shown at B is arranged so that it encloses some of the magnetic lines of force. The point at which maximum coupling will be secured depends upon the particular mode of propagation in the guide or cavity; the coupling will be maximum when the coupling device is in the most intense field.

Coupling can be varied by turning either the probe or loop through a 90-degree angle. When the probe is perpendicular to the elec-

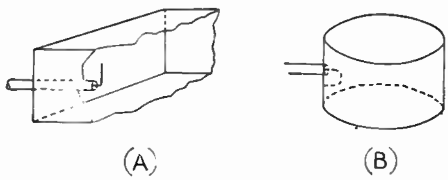


Fig. 15-28 — **Coupling to wave guides and resonators.**

tric lines the coupling will be minimum; similarly, when the plane of the loop is parallel to the magnetic lines the coupling will have its least possible value.

Amateur Microwave Technique

All the microwave bands allotted to amateurs have been used experimentally for communication purposes. Complete description of the equipment used is beyond the scope of this text, but reference is made to various articles which have appeared in *QST*, describing the gear devised by the amateur pioneers in this field.

For the experimentally-inclined, our microwave assignments represent a challenge to amateur ingenuity. Who can say but greater use of these frequencies will repeat past history, turning up propagation peculiarities and potential uses which will make these bands as coveted a region as our "communication frequencies" are considered today?

The first amateur microwave communication was carried on by Merchant and Harrison, W6BMS/2 and W2LGF, who assembled the gear shown in Fig. 15-29 in time to communicate with each other on November 15, 1945 — the date that the microwave bands were officially opened to amateur experimentation. They used two klystron tubes, one as a fre-

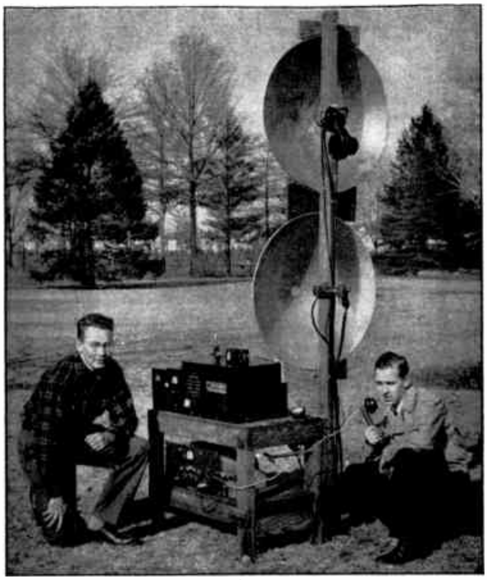


Fig. 15-29 — The first amateur microwave communication was accomplished by W6BMS (left) and W2LGF, who used two sets of similar equipment to open the 5300-Mc. amateur band on November 15, 1945, the date that the first microwave bands were released for amateur use.

quency-modulated transmitter oscillator, the other as a local oscillator for receiving. The latter worked in conjunction with a crystal mixer, into a 30-Mc. i.f. in the form of an FM receiver.

The 2300-Mc. amateur assignment was first used for communication by Koch and Floyd, W9WHM/2 and W6OJK/2, who used lighthouse tubes in simple transceivers, both of which are shown in Fig. 15-30. Their antenna systems used parabolic reflectors, one being made of wire screening attached to a wooden frame, and the other, also shown in the photograph, was simply an electric-heater assembly, with the microwave dipole substituted for the heater element.

Amateur communication on 10,000 Mc. was first accomplished by Atwater and McGregor, W2JN and W2RJM, who modified 723-A/B klystrons to permit their operation in the amateur band. They are shown, with one of their equipment set-ups, in Fig. 15-31. A somewhat similar arrangement was used by W4HPJ/3 and W6IFE/3 to extend the distance record, and has since been employed by W6IFE in opening the 3300-Mc. band to amateur use, except that the tube used in the latter instance was a 707-B with an external cavity.

The highest frequency ever used in amateur work is 21,000 Mc., first employed by Sharbaugh and Watters, W1NVL/2 and W9SAD/2, whose laboratory set-up is shown in Fig. 15-32. The r.f. generator, for transmitting and receiving, was a developmental tube designated as the Z-668, a velocity-modulated tube of the reflex type. Communication was carried on, two-way, over a distance of 800 feet.

A list of *QST* references, arranged according to the amateur band concerned, follows. It should be emphasized that the equipment described in these reports is experimental in nature. In most instances it represents only one of several ways in which microwave communication equipment might be built. The distances covered in the pioneering work just mentioned are not, for the most part, indicative of the

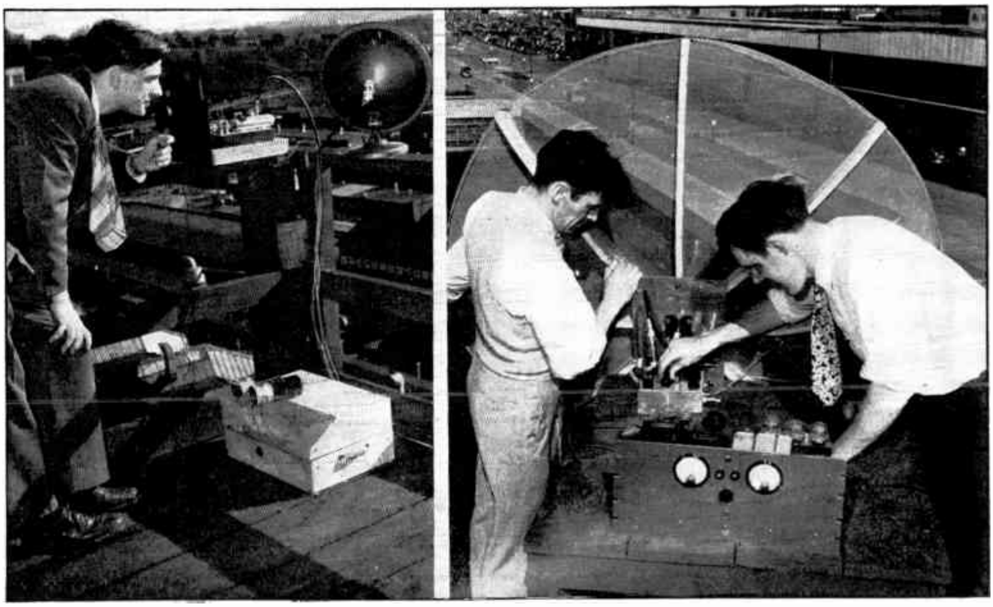


Fig. 15-30 — The 2300-Mc. band was first employed for amateur communication by W6OJK (left) and W9WHM (extreme right). Antenna systems employed a standard electric-heater unit and a handmade screen-lined parabola.

maximum working range, since exploration of the particular band in question was the end in view when the experiments were conducted, rather than the covering of any long distances.

Bibliography

1215 Mc. — "World Above 50 Mc." (W1BBM), May 1947 *QST*, page 136; also July 1947 *QST*, page 136. Sulzer and Ammerman, "An Oscillator for the 1215-Mc. Band," April 1948 *QST*, page 16.

2300 Mc. — Koch and Floyd, "CQ – 2400 Mc.," July 1946 *QST*, page 32. Also "World Above 50 Mc." (W6IFE), Aug. 1947 *QST*, page 128.

3300 Mc. — "World Above 50 Mc." (W6IFE), Aug. 1947 *QST*, page 128.

Fig. 15-31 — W2RJM (left) and W2JN, with one of the equipments used in pioneering work on 10,000 Mc.

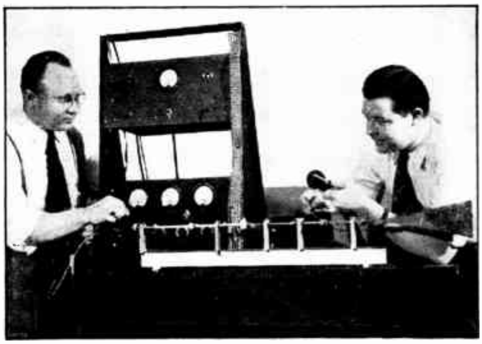


Fig. 15-32 — W1NVL (left) and W9SAD with the equipment used to work a distance of 800 feet on the highest frequency ever used for amateur communication — 21,000 Mc. Antenna systems employed a parabolic reflector at one end and a horn radiator at the other.

5250 Mc. — Merchant and Harrison, "Duplex 'Phone on 5300 Mc.," Jan. 1946 *QST*, page 19.

10,000 Mc. — McGregor and Atwater, "Dishing Out the Milliwatts on 10 KMc.," Feb. 1947 *QST*, page 58. Also "World Above 50 Mc." (W4HPJ/3, W6IFE/3), Sept. 1946 *QST*, page 152.

21,000 Mc. — Sharbaugh and Watters, "Our Best DX — 800 Feet!" Aug. 1946 *QST*, page 19.

Measuring Equipment

To comply with FCC regulations it is necessary that the amateur station be equipped to make a few relatively simple measurements. For example, the regulations require that means be available for checking the transmitter frequency to make sure that it is inside the band. This means must be independent of the frequency control of the transmitter itself; it is not enough to depend on, say, the calibration of a crystal in the crystal-controlled oscillator that drives the transmitter. In addition, it is necessary to make sure that the plate power input to the final stage of the transmitter does not exceed one kilowatt. The regulations also impose certain requirements with respect to plate-supply filtering, stability and purity of the transmitted signal, and depth of modulation in the case of 'phone transmission.

In many cases all these measurements can be made to a satisfactory degree of accuracy with no more auxiliary equipment than the regular station receiver. However, a better job usually can be done by building and calibrating some relatively simple test gear. Too, the progressive amateur is interested in instruments as an aid to better performance.

Frequency Measurement

Types of Equipment

Frequency-measuring equipment can be divided into two broad classes: oscillators of various types generating signals of known frequency that can be compared with the signal whose frequency is unknown, and adjustable resonant circuits.

Instruments in the first classification are the more accurate. Two types are commonly used by amateurs, the **secondary frequency standard** and the **heterodyne frequency meter.** The secondary frequency standard usually generates a frequency of 100 kc. and employs a circuit that is rich in harmonic output. As a result, it supplies a series of frequencies, all multiples of 100 kc., which provide accurate calibration points throughout the communications spectrum. The more elaborate instruments of this type include frequency dividers (multivibrators) to supply intermediate calibration points; a divisor commonly used is 10, thus signals are generated at intervals of 10 kc. when the fundamental frequency is 100 kc.

The conventional type of heterodyne frequency meter is simply a variable-frequency oscillator. The oscillator usually is designed to cover the lowest frequency band in which measurements are to be made; measurements then can be made in higher frequency bands by using the harmonic output of the oscillator. For example, when the oscillator is set to 3560 kc. its second harmonic is 7120 kc., its fourth harmonic is 14,240 kc., and so on. The proper frequency reading is determined by observing the fundamental frequency of the oscillator and then multiplying by the number of the harmonic that falls in the desired frequency range.

In both types of instruments — secondary standard and heterodyne meter — the inherent accuracy is a fixed percentage of the frequency at which the measurement is made. The secondary standard is usually the more accurate, since it can be made crystal-controlled with attendant high stability. However, it lacks the flexibility of the heterodyne meter in that it does not in itself provide a means for making measurements between adjacent harmonics of the oscillator or multivibrator. A third type of instrument uses a secondary standard in conjunction with a variable oscillator for interpolation. When these are combined in the "additive" frequency meter as described later, the result is a frequency meter that has essentially the accuracy of the secondary standard but has the direct measurement feature of the heterodyne meter.

Frequency-measuring equipment incorporating oscillators is used in conjunction with a regular receiver. The process of measurement consists of comparing the signal from the frequency meter with the signal whose frequency is to be measured. Nonoscillating types of frequency meters operate by absorbing some energy from the signal source under measurement, and in consequence are called "absorption" frequency meters. They are simply tuned circuits, adjustable over the desired frequency range, provided with some means for indicating when the energy in the circuit is maximum. Their accuracy is low compared with the oscillating types, but where approximate measurement is sufficient they have a number of desirable features.

Frequency Measurement with the Receiver

An ordinary receiver has the essential elements needed for frequency measurement. Its dial readings must be calibrated in terms of frequency, of course, before measurements can be made. Manufactured receivers are generally so calibrated; the accuracy of the calibration will vary with the receiver model, but if the receiver is well made and has good inherent stability, a bandspread dial calibration can be relied upon to within perhaps 0.2 per cent. For most accurate measurement, maximum response in the receiver should be determined by means of a carrier-operated tuning indicator (such as an S-meter), the receiver beat oscillator being turned off. If the receiver has a crystal filter, it should be set in a fairly "sharp" position to increase the accuracy.

When checking the frequency of your own transmitter, the receiving antenna should be disconnected so the signal will not overload or "block" the receiver. Also, the r.f. gain should be reduced as a further precaution against overloading. If the receiver still blocks without an antenna the frequency may be checked by turning off the power amplifier and tuning in the oscillator alone. It is difficult to avoid blocking under almost any conditions with a regenerative receiver, and so this type is not very suitable for checking the frequency of one's own transmitter.

WWV SCHEDULES

Standard radio and audio frequencies are broadcast continuously, day and night, from WWV, the station of the Central Radio Propagation Laboratory, National Bureau of Standards, Washington, D. C., on the following frequencies:

Mc.	*Power (kw.)*	*Audio Freq. (cycles)*
2.5	0.7	1 and 440
5.0	8.0	1 and 440
10.0	9.0	1, 440 and 4000
15.0	9.0	1, 440 and 4000
20.0	8.5	1, 440 and 4000
25.0	0.1	1, 440 and 4000
30.0	0.1	1 and 440
35.0	0.1	1

The 1-c.p.s. modulation is a 0.005-second pulse, the beginning of which marks the beginning of each second to an accuracy of one part in 1,000,000. The pulse is omitted on the 59th second of every minute.

The accuracy of the radio and audio frequencies is within one part in 50,000,000. The audio frequencies are interrupted at precisely one minute before each hour and each five minutes thereafter (59th minute, 4 minutes past hour, etc.); they are resumed in precisely one minute. During each silent interval the time (EST) is given in telegraphic code. A station announcement is given in voice on the hour and half hour.

THE SECONDARY FREQUENCY STANDARD

The most practical type of secondary standard for amateur use is a 100-kc. crystal oscillator. It is very simple to build and its harmonics will mark the edges of the amateur bands to a high degree of accuracy. A series of such "marker" signals at the band edges is all that is required, from the standpoint of making sure that the transmitter frequency is inside the band on which it is supposed to be working.

Manufacturers of 100-kc. crystals usually supply circuit information for their particular crystals. The circuit given in Fig. 16-1 is representative, and will generate usable harmonics up to 30 Mc. or so. The variable condenser, C_1, provides a means for adjusting the frequency to exactly 100 kc. Harmonic output is taken from the circuit through a small condenser, C_5. There are no particular constructional points to be observed in building such a unit. Power for the tube heater and plate may be taken from the supply in the receiver with which the unit is to be used. The plate voltage is not critical, but it is recommended that it be taken from a VR-150 regulator if the receiver is equipped with one.

Sufficient signal strength usually will be secured if a wire is run between the output terminal connected to C_5 and the antenna post on the receiver. At the lower frequencies a metallic connection may not be necessary.

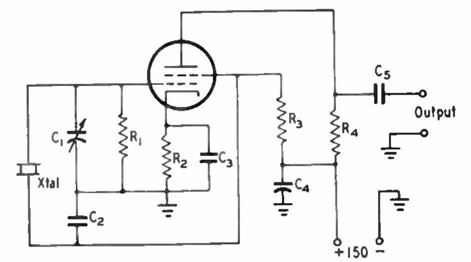


Fig. 16-1 — **Circuit for crystal-controlled frequency standard. Tubes such as the 6SK7, 6SH7, 6AU6, etc., are suitable.**

C_1 — 50-μμfd. variable.
C_2 — 150-μμfd. mica.
C_3 — 0.0022-μfd. mica.
C_4 — 0.01-μfd. paper.
C_5 — 22-μμfd. mica.
R_1 — 0.47 megohm, ½ watt.
R_2 — 1000 ohms, ½ watt.
R_3 — 0.1 megohm, ½ watt.
R_4 — 0.15 megohm, ½ watt.

Adjusting to Frequency

The frequency can be adjusted exactly to 100 kc. by making use of the WWV transmissions tabulated in this chapter. Select the frequency that gives a good signal at your location at the time of day most convenient. Tune in the WWV signal with the receiver b.f.o. off and wait for the period during which the modulation is absent. Then switch on the 100-kc. oscillator and adjust its frequency, by means of C_1, until its harmonic is in zero beat with WWV. The exact setting is easily found by observing the slow pulsation in background noise as the harmonic comes close to zero beat, and adjusting to where the pulsation disappears or occurs at a very slow rate. The pulsations can be observed even more readily by switching on the receiver's b.f.o., after approximate zero beat has been secured, and observing the rise and fall in intensity (not frequency) of the beat tone. For best results the WWV signal and the signal from the 100-kc. oscillator should be about the same strength. It is advisable not to try to set the 100-kc. oscillator when the WWV signal is modulated, since it is difficult to tell whether the harmonic is being adjusted to zero beat with the carrier or one of the sidebands.

"Marker" Frequencies

Identification of the 100-kc. harmonics is usually not difficult in or near the amateur bands because the normal activity in those bands will show which 100-kc. harmonics define the band limits. In other regions harmonics can be identified by counting them off from one whose frequency is known. The frequency of a given harmonic can often be identified by comparing it with a commercial or government station of known frequency operating in the vicinity. Alternatively, a "marker" crystal can be used. A favorite frequency for such a marker is 1000 kc. Harmonics of a 1000-kc. oscillator are easily identified on the average receiver because they are fairly widely spaced, and once the receiver setting for a multiple of 1000 kc. is determined it is an easy matter to count off the 100-kc. points between. Other marker frequencies can of course be used — for example, a frequency near 2000 kc., which is in the range of crystals available for amateur use. The circuit given in Fig. 16-1 will work satisfactorily with such crystals, so the marker points can be determined simply by inserting a suitable crystal.

THE HETERODYNE FREQUENCY METER

The basis of the heterodyne frequency meter is a completely-shielded oscillator with a precise frequency calibration. The oscillator must be so designed and constructed that it can be accurately calibrated and will retain its calibration over long periods of time.

The oscillator used in the frequency meter must be very stable. Mechanical considerations are most important in its construction. No matter how good the instrument may be electrically, its accuracy cannot be depended upon if the mechanical construction is flimsy. Inherent frequency stability can be improved by avoiding the use of phenolic compounds and thermoplastics (bakelite, polystyrene, etc.) in the oscillator circuit, employing only high-grade ceramics instead. Plug-in coils ordinarily are not acceptable; instead, a solidly-built and firmly-mounted tuned circuit should be permanently installed. The oscillator panel and chassis should be as rigid as possible.

To be usable over a wide frequency range the heterodyne frequency meter must have strong harmonic output. A suitable circuit, including a harmonic amplifier, is shown in Fig. 16-2. The mechanical construction should parallel that of the VFOs shown in Chapter Six. In the oscillator circuit, an adjustable padding condenser, C_2, is provided so that the tuning range can be set to cover whichever band is selected for the fundamental frequency. In addition, it may be necessary to adjust the coil inductance slightly in order to make the range cover as much as possible of the tuning dial.

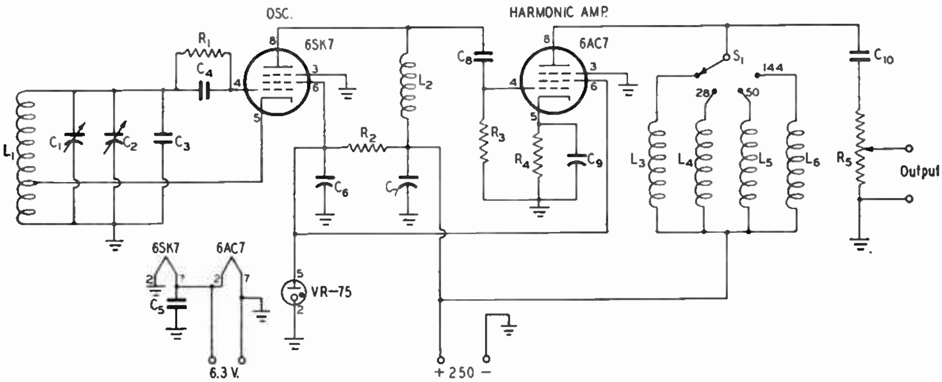


Fig. 16-2 — Heterodyne frequency meter with harmonic amplifier.

C_1 — 100-μμfd. variable (tuning).
C_2 — 100-μμfd. variable (band-set).
C_3 — 220-μμfd. silver mica (padder).
C_4, C_8, C_{10} — 100-μμfd. mica.
C_5, C_6, C_7, C_9 — 0.01-μfd. paper.
R_1, R_3 — 0.47 megohm, ½ watt.
R_2 — 10,000 ohms, 1 watt.
R_4 — 330 ohms, 1 watt.
R_5 — 25,000-ohm potentiometer.
L_1 — For 3500–4000 kc. fundamental: 18 turns No. 18 on 1-inch form, length 1½ inches. Cathode tap 5 turns from ground end.
For 1750–2000 kc. fundamental: 36 turns No. 20 d.c.c. close-wound on 1-inch form. Cathode tap 10 turns from ground end.
L_2, L_3 — 2.5-mh. r.f. choke.
L_4 — 24 turns No. 18 enam. close-wound on ¼-inch form.
L_5 — 11 turns No. 18 enam. close-wound on ¼-inch form.
L_6 — 2 turns No. 16 spaced ½ inch, diameter ¼ inch.
S_1 — 4-position 1-pole ceramic wafer switch.

Although the oscillator alone will give satisfactory output in the lower-frequency amateur bands, better results at 28 Mc. and higher are obtained by using the 6AC7 harmonic amplifier. The 6AC7 plate circuit is broadly tuned by means of switched coils resonating, with the circuit capacitances, at 144, 50 and 28 Mc. A radio-frequency choke is connected to the fourth switch position; this gives ample signal strength at 14 Mc. and lower frequencies. Potentiometer R_5 in the output circuit makes it possible to reduce the strength of the signal from the frequency meter to the value desired for measurement purposes.

The various amateur bands are covered by the following harmonics: 3.5–4 Mc., fundamental; 7–7.3 Mc., 2nd harmonic; 14–14.4 Mc., 4th; 26.96–27.23 Mc., 7th; 28–29.7 Mc., 8th; 50–54 Mc., 14th; 144–148 Mc., 40th. At lower frequencies a short length of wire connected to the output terminal will give ample signal strength under average conditions, but in the v.h.f. range closer coupling — such as running the wire in close proximity to the receiving antenna lead, or actually connecting it to the antenna post through a small fixed condenser — may be necessary to get a good signal.

Calibration

The heterodyne frequency meter may be calibrated against the harmonics of a 100-kc. secondary standard of the type described in the preceding section, using a receiver as an auxiliary. For example, suppose the oscillator fundamental range is 3.5–4 Mc. Then if the receiver is adjusted to pick up the fifth harmonic of the oscillator (17.5 to 20 Mc.) and the harmonic is beat against 100-kc. points from the crystal oscillator in that range, 100-kc.

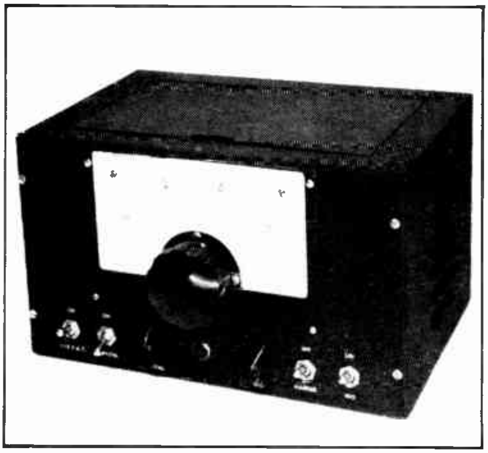


Fig. 16-3 — Additive frequency meter with self-contained power supply. The small knobs are for correction of drift so that both the 100-kc. crystal oscillator and VFO can be set to exact frequency. Dial calibration is in 1000-cycle intervals. This unit can be used for high-accuracy frequency measurement at all frequencies from 100 kc. through 30 Mc.

intervals on the fifth harmonic will give 20-kc. intervals on the fundamental. With a straight-line capacitance condenser at C_1, the relationship between dial divisions and frequency is almost linear, and marking off the dial at the proper intervals between actual calibration points will result in a calibration of sufficient accuracy.

INTERPOLATION-TYPE FREQUENCY METER

By using a variable-frequency oscillator of restricted tuning range to interpolate between the harmonics generated by a 100-kc. crystal standard, it becomes possible to measure frequency with an accuracy that is more than adequate for all practical purposes. In the frequency meter shown in Figs. 16-3 to 16-6, inclusive, this interpolation is accomplished by modulating the harmonic output of the 100-kc. oscillator with the output of a 100–150-kc. variable oscillator. As in ordinary telephony, the modulation process sets up side frequencies that add algebraically to each harmonic, hence the name "additive frequency meter." The sidebands appear as signals of adjustable frequency between the 100-kc. harmonics.

To cover a 100-kc. range, the interpolation oscillator need cover only an actual tuning range of 50 kc. This is because both sum and difference frequencies appear. For example, if the VFO is set at 100 kc., this frequency will add to and subtract from each harmonic of the crystal oscillator. Thus the crystal harmonic at 6900 kc., when modulated by 100 kc., will produce side frequencies at 7000 kc. and 6800 kc.; likewise, the crystal harmonic at 7200 kc. will have side frequencies at 7300 and 7100 kc. If the VFO is set to 150 kc., the same crystal harmonics will have side frequencies at 7050 and 6750 kc., and at 7350 and 7050 kc., respectively. In the latter case the upper side frequency of the 6800-kc. harmonic coincides with the lower side frequency of the 7200-kc. harmonic, both being at 7050 kc. Hence the same VFO signal, in tuning from 100 to 150 kc., covers the range from 7000 to 7050 kc., and from 7100 to 7050 kc., simultaneously. This occurs between each pair of 100-kc. crystal harmonics throughout the spectrum. Since the side frequencies move in opposite directions when the tuning of the VFO is varied, the interpolation scale is calibrated to read from 0–50 kc. (corresponding to varying the actual VFO frequency from 100 to 150 kc.) in one direction, and from 50–100 kc. in the opposite direction.

The circuit diagram of the instrument is shown in Fig. 16-4. A double triode is used as a combination VFO-amplifier, the amplifier being of the cathode-follower type to provide good isolation. The output of the amplifier goes through a low-pass filter (C_{13}, C_{14}, C_{15}, L_2, L_3 and R_7) to prevent oscillator harmonics

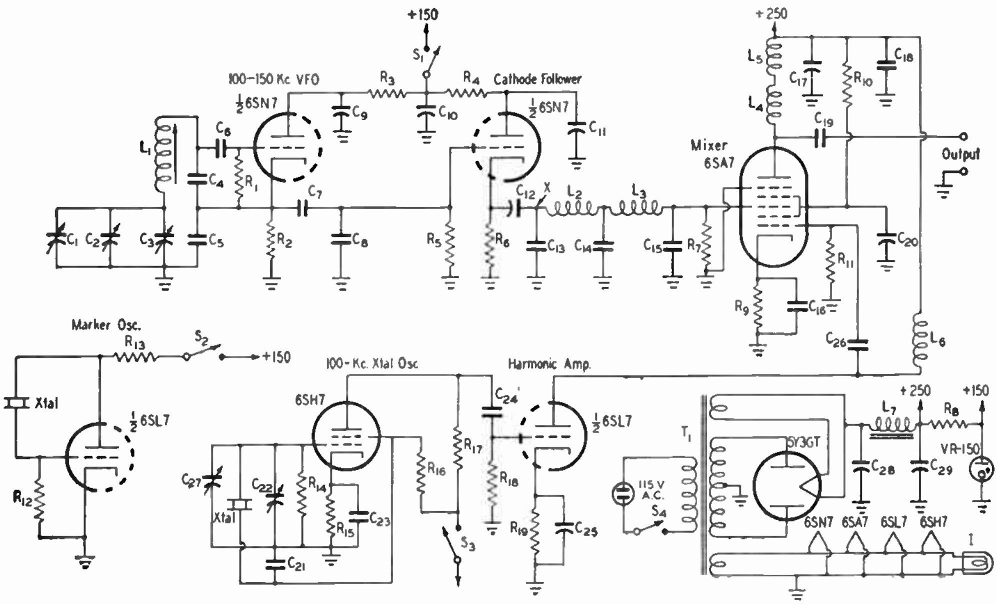


Fig. 16-4 — Circuit diagram of the additive frequency meter.

C_1 — 25-μμfd. variable (Millen 20025) (drift corrector).
C_2 — 100-μμfd. variable (Millen 26100) (padder).
C_3 — 250-μμfd. variable (National SEH-250) (tuning).
C_4, C_5, C_{23} — 0.0022-μfd. mica.
C_6, C_7, C_{19} — 470-μμfd. mica.
C_8, C_{16}, C_{18} — 0.001-μfd. mica.
C_9, C_{11}, C_{20} — 0.1-μfd. paper.
C_{10}, C_{12}, C_{17}, C_{25} — 0.01-μfd. paper.
C_{13}, C_{15} — 680-μμfd. mica.
C_{14} — 1360-μμfd. mica (two 680-μμfd. units in parallel).
C_{21} — 150-μμfd. mica.
C_{22} — 50-μμfd. variable (Millen 26050).
C_{24} — 22-μμfd. mica.
C_{26} — 100-μμfd. mica.
C_{27} — 15-μμfd. variable (Millen 20015).
C_{28}, C_{29} — 8-μfd. electrolytic, 450 volts.
R_1 — 47,000 ohms, ½ watt.
R_2, R_{10} — 22,000 ohms, 1 watt.
R_3 — 3300 ohms, ½ watt.
R_4 — 2200 ohms, ½ watt.
R_5, R_{12}, R_{14}, R_{18} — 0.47 megohm, ½ watt.
R_6, R_{15}, R_{19} — 1000 ohms, ½ watt.
R_7 — 1500 ohms, ½ watt.
R_8 — 2500 ohms, 10 watts.
R_9 — 220 ohms, ½ watt.
R_{11} — 0.22 megohm, ½ watt.
R_{13} — 0.1 megohm, 1 watt.
R_{16} — 0.1 megohm, ½ watt.
R_{17} — 0.15 megohm, ½ watt.
L_1 — Variable from app. 8 to 11 mh. (Millen 65000-35).
L_2, L_3 — 2.5-mh. r.f. choke (National R-50).
L_4 — 10 μh. (National R-60).
L_5 — 100 μh. (National R-33).
L_6 — 7 μh. (Ohmite Z-50).
L_7 — 40-ma. filter choke.
I — Pilot-lamp assembly.
S_1, S_2, S_3, S_4 — S.p.s.t. toggle.
T_1 — Power transformer, 275 each side c.t. at 50 ma.; 6.3 v. at 2.5 amp.; 5 v. at 2 amp. (Thordarson T22R30).

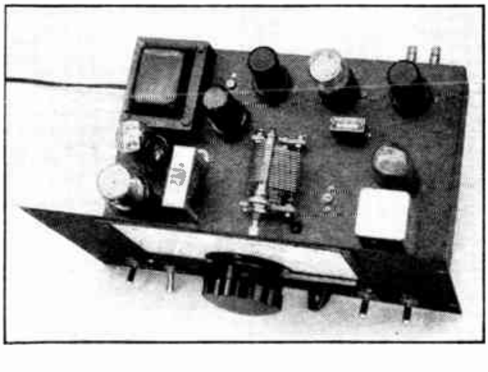
from being applied to the 6SA7 modulator or mixer tube. The output of the 6SH7 100-kc. crystal oscillator is fed through a harmonic amplifier (one 6SL7 section) before also being applied to the mixer tube, the purpose being to level off the harmonic strength throughout the spectrum as much as possible. The plate circuit of the mixer is likewise adjusted so that the output signal is as uniform in strength as possible up to 30 Mc. The spare triode section of the 6SL7 is used as an auxiliary crystal oscillator so that a marker crystal can be used for identification of the 100-kc. crystal harmonics.

♦

Fig. 16-5 — Chassis view of the additive frequency meter. Immediately in front of the power transformer are the rectifier and voltage-regulator tubes. The 100-kc. crystal, mounted in a metal-tube shell (James Knights), is just to the right of the power transformer. The tubes along the rear edge, from left to right, are the 6SH7, 6SL7, and 6SA7. The marker crystal is immediately in front of the 6SL7. The VFO coil is at the lower right, with the 6SN7 just behind it. The shaft for the oscillator padder projects through the chassis to the right of the tuning condenser.

Calibration

To set up the instrument, it is necessary first to adjust the VFO range exactly to 100-



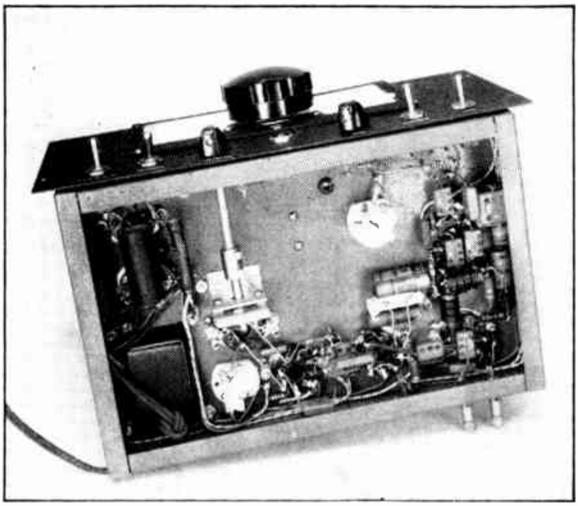


Fig. 16-6 — Bottom view of the frequency meter. Parts can be identified by reference to the tubes with which they are associated (see top view).

150 kc. For this purpose the 6SL7 and 6SA7 should be out of their sockets. On any receiver capable of tuning to 600 kc., tune in the 6th harmonic of the 100-kc. crystal oscillator. Connect a wire from point X to the antenna post of the receiver. Turn the VFO condenser over its whole range and note the number of harmonics heard at 600 kc., C_2 being at about 75 per cent of full scale. Adjust L_1, and C_2 if necessary, until there are just three such harmonics, one at each end of the scale and one between. This adjusts the oscillator to the proper range, by making the 4th harmonic of the high end and the 6th harmonic of the low end fall at 600 kc.

After noting the strength of the oscillator harmonics, shut off the 100-kc. crystal oscillator and move the receiver antenna connection from X to the No. 3 grid connection (output of the harmonic filter) on the 6SA7 socket. It should be impossible to hear any harmonic output from the oscillator when the tuning is varied. Then insert the 6SA7 in its socket, allow it to warm up, and again tune the VFO over its range. If harmonics now become audible the oscillator signal is too strong. It may be reduced by increasing the capacitance at C_8 as much as is necessary to make the harmonics disappear.

Calibration is best carried out in a series of steps. Remove the 6SA7 and 6SL7, connect the receiver antenna post to point X, and tune in the 2000-kc. harmonic from the 100-kc. crystal oscillator. Set the VFO at 100 kc., and bring its harmonic to zero beat with the crystal harmonic. Mark this point "0" on the dial. Then tune the receiver to the 21st crystal harmonic (2100 kc.) and slowly tune the VFO higher in frequency until its harmonic is at zero beat with the crystal harmonic. At this point the 20th harmonic of the VFO coincides with the 21st harmonic of the crystal, and so the VFO frequency is 2100/20 = 105 kc. Mark this point "5" on the scale, move the receiver to 2200 kc., and increase the VFO frequency until its 20th harmonic coincides with 2200 kc., giving the 10-kc. point. Continue until the scale is calibrated at each 5-kc. point up to 50 kc.

The next step is to calibrate at 2-kc. intervals, and for this purpose it is necessary to increase the strength of the harmonics. The marker oscillator can be used as an amplifier, by removing the crystal and making the connections shown in Fig. 16-7A. Clip leads are satisfactory. It is necessary to replace the 6SL7, but do not put the 6SA7 in its socket. Tune in the 5000-kc. harmonic of the 100-kc. crystal oscillator, set the VFO to 100 kc. by beating its 50th harmonic with the 5000-kc. harmonic of the crystal, and proceed up through the spectrum one 100-kc. point at a time, using the same procedure as before. The VFO harmonics will tune quite rapidly, and the previously-determined 5-kc. marks will ensure that the calibration points do not get out of proper order.

The impromptu harmonic amplifier alone will not usually give enough output to repeat this process with the 100th harmonic, by means of which 1-kc. points are obtained. The necessary harmonics can be generated by using a crystal rectifier as shown in Fig. 16-7B. In this case the lead from the receiver antenna should be brought near, but not connected to, the harmonic amplifier. The crystal acts as a mixer and introduces many secondary beats, but if the coupling to the receiver is loose enough the desired harmonics will be the strongest and can easily be identified, particularly since the 2-kc. points already plotted will practically show where they should fall. There should also be no trouble in hearing the 100-kc. crystal harmonics from 10 to 15 Mc. if the receiver antenna lead is near the crystal oscillator. The calibration points should be plotted on the scale as accurately as possible.

By use of the drift-corrector condensers the accuracy of the instrument is practically the

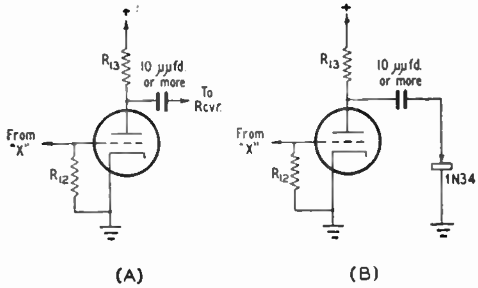


Fig. 16-7 — Temporary connections for amplifying VFO harmonics when calibrating. The marker-oscillator tube is used with the crystal removed.

accuracy with which the dial can be read. Interpolation to 100 cycles is readily possible. The crystal-oscillator frequency can be checked against WWV and reset when accurate measurements are to be made. The VFO is easily corrected by setting the dial to the 50-kc. point and adjusting the drift-corrector condenser to bring the two side frequencies into exact zero beat. Without drift correction the instrument is reliable to the nearest kilocycle, with average construction and good components, at frequencies as high as 30 megacycles.

(A complete description of this system is given in May, 1949, *QST*.)

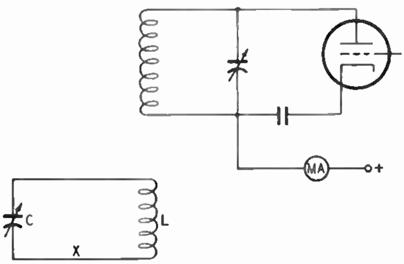


Fig. 16-8 — **Absorption frequency meter and a typical application. The meter consists simply of the resonant circuit *LC*. When coupled to an amplifier or oscillator the tube plate current will rise when the frequency meter is tuned to resonance. The frequency may then be read from a calibrated condenser dial. Suitable constants for *L* and *C* may be taken from Fig. 16-10. A flashlight lamp may be connected in series at *X* to give a visual indication, but it decreases the selectivity of the instrument and makes it necessary to use rather close coupling to the circuit being measured.**

ABSORPTION FREQUENCY METERS

The simplest possible frequency-measuring device is a resonant circuit, tunable over the desired frequency range and having its tuning dial calibrated in terms of frequency. Such a frequency meter operates by extracting a small amount of energy from the oscillating circuit to be measured, the frequency being determined by the tuning setting at which the energy absorption is maximum.

This method is not capable of as high accuracy as the heterodyne methods for two reasons: First, the resonance indication is relatively "broad" as compared with the zero beat of a heterodyne; second, the necessarily close coupling between the frequency meter and the circuit being measured causes some detuning in both circuits, with the result that the calibration of the frequency-meter circuit depends to some degree on the coupling to the circuit being measured. Nevertheless, an absorption wavemeter is a highly useful instrument in the amateur station. It requires no power supply for its operation, which is a convenience. It also eliminates the confusion that sometimes arises because of the large number of harmonic responses that occur in making measurements by heterodyne methods: a simple tuned circuit will respond to only one frequency. This is helpful, for example, in determining the actual output frequency of a frequency multiplier in the transmitter, and eliminates the possibility that the multiplier can be tuned to the wrong harmonic.

When an absorption meter is used for checking a transmitter, the plate current of the tube connected to the circuit being checked can provide the necessary resonance indication. When the frequency meter is tuned through resonance the plate current will rise, and if the frequency meter is loosely coupled to the tank circuit the plate current will simply give a slight upward flicker as the meter is tuned through resonance. The greatest accuracy is secured when the loosest possible coupling is used.

A receiver oscillator may be checked by tuning in a steady signal and heterodyning it to give a beat note as in ordinary c.w. reception. When the frequency meter is coupled to the oscillator coil and tuned through resonance the beat note will change. Again, the coupling should be made loose enough so that a just-perceptible change in beat note is observed when the meter is tuned through resonance.

An approximate calibration may be obtained by comparison with a calibrated receiver. The usual receiver dial calibration is sufficiently accurate. A simple oscillator circuit covering the same range as the frequency meter will be useful in calibration. Set the receiver to a given frequency, tune the oscillator to zero beat at the same frequency, and adjust the frequency meter to resonance with the os-

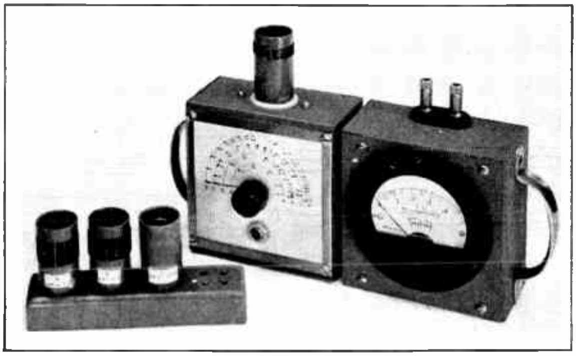


Fig. 16-9 — **A sensitive absorption-type frequency meter with a crystal-detector rectifier and a d.c.-milliammeter indicating circuit. The meter is housed in a separate compartment so that it may be used with other measuring devices. The cabinet and front cover are drilled and tapped to accommodate the mounting screws for a large-size chart frame: frequency calibrations are marked on cardboard held in place by the chart frame. A short strip of wood, drilled to match the coil-form prongs, is used as a rack for the coils. Meter-box connections are shown in Fig. 16-20.**

cillator as described above. This gives one calibration point. When a sufficient number of such points has been obtained a graph may be drawn to show frequency *vs.* dial settings on the frequency meter.

A Sensitive Absorption Frequency Meter

Figs. 16-9 to 16-11, inclusive, show an absorption frequency meter or "wavemeter" with a crystal-detector/milliammeter resonance indicator that provides a relatively high degree of sensitivity. As shown in Fig. 16-10, a resonant circuit is connected in series with a crystal detector and a 0–1 milliammeter (a microammeter can be substituted for still greater sensitivity). The tank coil, L_1, serves as the pick-up coil, and the crystal is tapped down on the inductance in order to improve the sensitivity and selectivity of the meter. Plug-in coils are provided so that the unit covers a frequency range from about 1 megacycle to 165 megacycles. Any type of fixed crystal detector may be used, but the v.h.f. types are recommended. The meter box shown at the right in Fig. 16-9 is the same unit that is used with the volt-ohm-milliammeter described later in this chapter.

The frequency meter is housed in a 2 × 4 × 4-inch metal box, the milliammeter being mounted in a separate box of the same size. The coil socket is on the top near the front edge, with the tuning condenser just below it inside the case. This arrangement keeps the tuned-circuit leads short. A headphone jack is provided for monitoring 'phone transmissions. The unit may be calibrated as described in the preceding section.

A two- or three-foot antenna rod may be added to the unit to permit using the instrument for field-strength measurements. The antenna should be connected to the top end of the tank coil, L_1. The rod antenna may be undesirable when the frequencies of individual simultaneously-operating circuits are to be checked — as in the case of a multistage transmitter with frequency multipliers — because the antenna increases the sensitivity to such an extent that it may be difficult to identify the output of a particular circuit. It may be convenient to interconnect the two units by means of a length of lamp cord or coaxial cable of any reasonable length (up to several hundred feet) when the meter is being used as a field-strength measuring device.

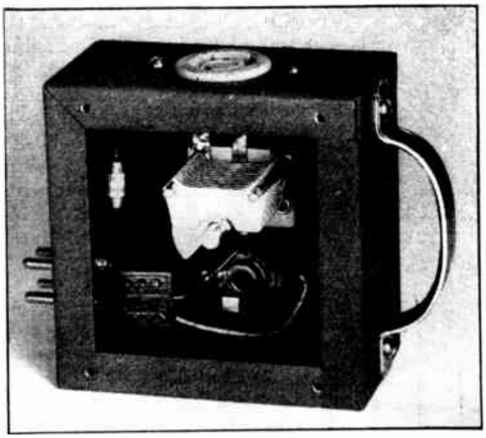


Fig. 16-11 — A rear view of the absorption-type frequency meter. The crystal is wired between the connector plug at the left and the coil socket at the top. The meter by-pass condenser is mounted between the plug and the grounded side of the 'phone jack. The variable-condenser terminals are connected directly to the coil socket.

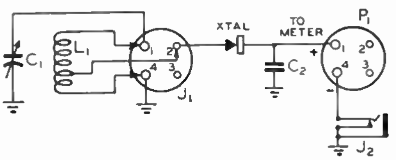


Fig. 16-10 — Circuit diagram of the absorption-type frequency meter.

C_1 — 140-μμfd. variable (Millen 22140).
C_2 — 0.0015-μfd. midget mica.
L_1 — 1.22–4.0 Mc.: 70 turns No. 32 enameled wire, 1-inch diam., 5/8 inch long. Tap 12½ turns from grounded end.
— 4.0–13.5 Mc.: 20 turns No. 20 enameled wire, 1-inch diam., 9/16 inch long. Tap 4½ turns from grounded end.
— 13.2–44.0 Mc.: 5 turns No. 20 enameled wire, 1-inch diam., 5/16 inch long. Tap 1½ turns from grounded end.
— 39.8–165 Mc.: Hairpin loop of No. 14 wire, ½-inch spacing, 2 inches long (total length including ends which fit down into the coil-form prongs). Tap 1⅝ inches from grounded end.
All four coils wound on Millen 45004 coil forms.
J_1 — 4-prong tube socket.
J_2 — Closed-circuit jack.
P_1 — 4-prong male plug.
XTAL — Type 1N34.

In addition to the uses mentioned in the preceding section, a meter of this type may be used for final adjustment of neutralization in r.f. amplifiers. For this purpose it may be loosely coupled to the plate tank coil. Alternatively, L_1 may be removed and the final-amplifier link output terminals connected to Prongs 2 and 4 in the coil socket. The latter method tends to ensure that the pick-up is from the final tank coil only.

● LECHER WIRES

At very-high and ultrahigh frequencies it is possible to determine frequency by actually measuring the length of the waves generated. The measurement is made by observing standing waves on a two-wire parallel transmission line or **Lecher wires.** Such a line shows pronounced resonance effects, and it is possible to determine quite accurately the current loops (points of maximum current). The physical distance between two consecutive current loops is equal to one-half wavelength. Thus the wavelength can be read directly in meters (39.37 inches = 1 meter; 0.3937 inch = 1 cm.), or in centimeters for the very-short wavelengths.

The Lecher-wire line should be at least a wavelength long — that is, 7 feet or more on 144 Mc. — and should be entirely air-insulated except where it is supported at the ends. It may be made of copper tubing or of wires stretched tightly. The spacing between wires should not exceed about 2 per cent of the shortest wavelength to be measured. The positions of the current loops are found by means of a "shorting bar," which is simply a metal strip or knife edge which can be slid along the line to vary its effective length. The system can be used more conveniently and with greater accuracy if it is built up in permanent fashion and provided with a shorting bar maintained at right angles to the wires (Fig. 16-12). The support may consist of two pieces of "1-by-2" pine fastened together with wood screws to form a "T"-girder, this arrangement being used to minimize bending of the wood when the wires are tightened. A slider holds the shorting bar and acts as a guide to keep the wire spacing constant.

For measuring lengths in the metric system used for wavelength, the supporting beam may be marked off in decimeter (10-centimeter) units. A 10-centimeter transparent scale (obtainable at 5 & 10 cent stores) may be cemented to the slider, extending out from the front, so that readings can be taken to the nearest millimeter. The difference between any two readings gives the half-wavelength directly.

Making Measurements

Let us suppose the frequency of a transmitter is to be measured. A convenient and fairly sensitive indicator can be made by soldering the ends of a one-turn loop of wire, of about the same diameter as the transmitter tank coil, to a low-current flashlight bulb, then coupling the loop to the tank coil to give a moderately bright glow. A coupling loop should be connected to the ends of the Lecher wires and brought near the tank coil, as shown in Fig. 16-13. Then the shorting bar should be slid along the wires outward from the transmitter until the lamp gives a sharp dip in brightness. This point should be marked and the shorting bar moved out until a second dip is obtained. The distance between the two points will be equal to half the wavelength. If the measurement is made in inches, the frequency will be

$$F_{Mc.} = \frac{5905}{length\ (inches)}.$$

If the length is measured in meters,

$$F_{Mc.} = \frac{150}{length\ (meters)}.$$

In checking a superregenerative receiver, the Lecher wires may be similarly coupled to the receiver coil. In this case the resonance indication may be obtained by setting the receiver just to the point where the hiss is obtained, then as the bar is slid along the wires a spot will be found where the receiver goes out of oscillation. The distance between two such spots is equal to a half-wavelength.

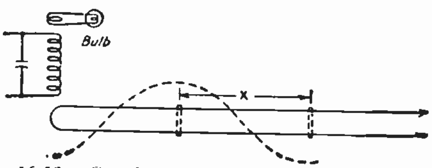


Fig. 16-13 — Coupling a Lecher wire system to a transmitter tank coil. Typical standing-wave distribution is shown by the dashed line. The distance X between the positions of the shorting bar at the current loops equals one-half wavelength.

The most accurate readings result when the loosest possible coupling is used between the line and the tank coil. After taking a preliminary reading to find the regions along the line in which resonance occurs, loosen the coupling until the indications are just discernible and repeat the measurement. As the coupling is loosened the resonance points will become sharper, which is a further aid to accurate determination of the wavelength.

The shorting bar must be kept at right angles to the two wires. A sharp edge on the bar is desirable, since it not only helps make good contact but also definitely locates the *point* of contact.

The accuracy with which frequency can be measured by such a system depends principally upon the technique of measurement. Careful measurement of the exact distance between two current loops is essential. An accurate standard of length is necessary — a good steel tape, for instance — for all but rough measurements.

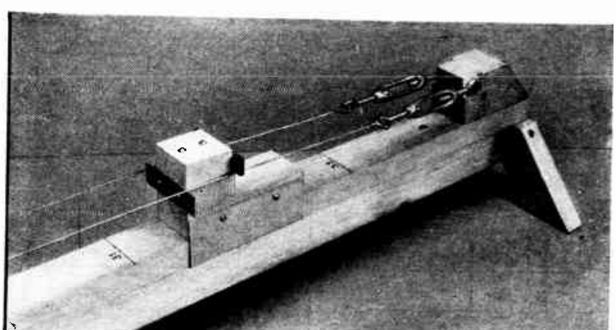


Fig. 16-12 — One end of a typical Lecher wire system. The feet at each end keep the assembly from tipping over when in use. The wire is No. 16 bare solid-copper antenna wire (hard-drawn). The turnbuckles are held in place by a 3/16 × 2-inch bolt through the anchor block. The other end of the line, the one connected to the pick-up loop, should be insulated.

Signal Monitoring

Every amateur should make provision for checking the quality of his transmitter's output. This requires that some means be available in the station for reducing the strength of the signal from the transmitter to the point where its characteristics can be examined without danger of false indications from overloading the receiving equipment.

The simplest method of checking the quality of c.w. transmissions is to use the regular station receiver. If the receiver is a superheterodyne the process may simply be that of reducing the r.f. gain to minimum and tuning to the transmitter frequency. If distant signals are stable and have "pure-d.c." tone in normal reception, then the local transmitter should, too, when the receiver gain is reduced to the point where the receiver does not overload. If the signal is too strong with the r.f. gain "off," shorting the receiver antenna input terminals may reduce it to suitable proportions, or the mixer circuit in the receiver may be temporarily detuned to arrive at the same desired result.

An alternative method is to set the receiver on the next lower-frequency band than the one in use, then tune the receiver so that the second harmonic of its oscillator beats with the transmitter signal to produce the intermediate frequency. Higher-order harmonics also may be used for this purpose. With this harmonic method there is ordinarily no danger that the receiver will overload, because the r.f. and mixer tuned circuits are so far from resonance with the transmitter frequency. The setting of the tuning dial bears no direct relation to the transmitter frequency under these conditions, since the oscillator harmonic must maintain a constant difference with the transmitter to produce the i.f. beat.

A 'phone signal may be monitored in the same way, provided a headset is used for reception. Use of a loudspeaker is not usually practicable because the sound output feeds back to the microphone and causes howling. A crystal detector and headset may also be used for the same purpose, as described in preceding sections. In monitoring a 'phone signal the best plan is to have another person speak into the microphone rather than to listen to one's own voice. It is difficult to judge quality when speaking and listening at the same time.

MODULATION MONITOR

Fig. 16-14 is the circuit of a 'phone monitor that can be used both for aural checking and for measuring modulation percentage. When a small r.f. voltage is applied to the input circuit it is rectified by the crystal. With switch S_1 in the "r.f." position the average value of the rectified current is measured by the 0–1 milliammeter, *MA*. With the switch in the "a.f." position, the audio modulation on the signal is transferred through T_1 to a second rectifier. The average value of the rectified audio is again read by the milliammeter. The circuit constants are chosen so that if the input is adjusted to make the meter read full scale on r.f., the a.f. meter readings will be directly proportional to percentage of modulation (for voice modulation), 100 per cent modulation being represented by a current of 1 milliampere. Switch S_2 provides for reversing the "polarity" of the modulation, giving a qualitative indication of the up- and down-peaks. A headphone jack, J_1, is provided for listening to the quality of the modulation. (The percentage modulation cannot be read with 'phones plugged into J_1, so the 'phones must be removed when readings are to be taken.)

In constructing such an instrument, care should be used to prevent r.f. pick-up in the audio rectifier circuit. This can be checked by testing the instrument on an unmodulated carrier (which must be substantially hum-free); with a full-scale reading when S_1 is in the "r.f." position, the meter should read zero when S_1 is switched to "a.f." The values of resistors R_1 and R_2 are critical and should be within plus or minus 5 per cent of the recommended values.

A sample of the modulated carrier may be coupled into the instrument through a one-turn link and a length of Twin-Lead, the link being placed within a few inches of the final tank circuit of the transmitter. The coupling between the link and final tank coil must be adjusted to give a full-scale r.f. reading, after

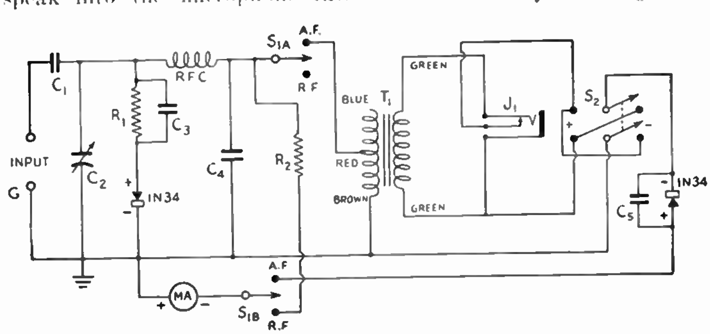


Fig. 16-14 — **Circuit of direct-reading modulation meter.**
C_1, C_4 — 1000-μμfd. ceramic.
C_2 — 100-μμfd. variable midget.
C_3 — 12-μμfd. mica.
C_5 — 470-μμfd. mica.
R_1 — 1100 ohms, 5%, 1 watt.
R_2 — 16,000 ohms, 5%, 1 watt.
J_1 — Closed-circuit jack.
MA — 0–1 ma., 100 ohms.
RFC — 20 μh.
S_{1A-B}, S_2 — D.p.d.t. toggle.
T_1 — Push-pull interstage transformer, 1:1 ratio.

C_2 has been set for maximum reading. Alternatively, a coil that will resonate with C_2 at the operating frequency may be connected to the input terminals and the instrument located so that a suitable full-scale reading will be obtained.

Besides indicating modulation percentage, the instrument will show carrier shift (as shown by a change in the reading, when modulating, with S_1 in the "r.f." position) and thus detect nonlinearity in the modulated amplifier.

Measurement of Current, Voltage and Resistance

D.C. Instruments

D.c. ammeters and voltmeters are basically identical instruments, the difference being in the method of connection. An ammeter is connected in series with the circuit and measures the current flow. A voltmeter indicates the current through a high resistance connected across the source to be measured; its calibration is in terms of the voltage drop in the resistance or multiplier.

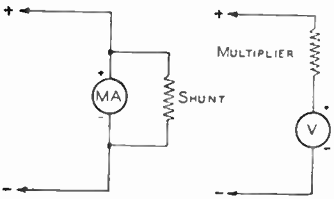


Fig. 16-15 — **How voltmeter multipliers and milliammeter shunts are connected to extend the range of a d.c. meter.**

If a single instrument must be used for measuring widely-different values of current or voltage, it is advisable to purchase one that will read, at about 75 per cent of full scale, the *smallest* value of current or voltage to be measured. Small currents cannot be read with any degree of precision on a high-scale instrument, but the range of a low-scale instrument can be extended as desired to take care of larger values. The ranges can be extended by the use of external resistors, connected in series with the instrument in the case of a voltmeter, and in parallel or "shunt" in the case of an ammeter. Fig. 16-15 shows at the left the manner in which a shunt is connected to extend the range of an ammeter and at the right the connection of a voltmeter multiplier.

To calculate the value of a shunt or multiplier it is necessary to know the internal resistance of the meter itself. If it is desired to extend the range of a voltmeter, the value of resistance which must be added in series is given by the formula

$$R = R_m (n - 1)$$

where R is the multiplier resistance, R_m the resistance of the voltmeter, and n the scale multiplication factor. For example, if the range of a 10-volt meter is to be extended to 1000 volts, n is equal to 1000/10 or 100.

If a milliammeter is to be used as a voltmeter, the value of series resistance can be found by Ohm's Law:

$$R = \frac{1000E}{I}$$

where E is the desired full-scale voltage and I the full-scale reading of the instrument in milliamperes.

To increase the current range of a milliammeter, the resistance of the shunt is

$$R = \frac{R_m}{n - 1}$$

where the symbols have the same meanings as above.

Homemade milliammeter shunts can be constructed from any of the various special kinds of resistance wire, or from ordinary copper magnet wire if no resistance wire is available. The Copper Wire Table in Chapter Twenty-Four gives the resistance per 1000 feet for various sizes of copper wire. After computing the resistance required, determine the smallest wire size that will carry the full-scale current (at 250 circular mils per ampere). Measure off enough wire (pulled tight but not stretched) to provide the required resistance. Accuracy can be checked by causing enough current to flow through the meter to make it read full scale without the shunt; connecting the shunt should then give the correct reading on the new full-scale range.

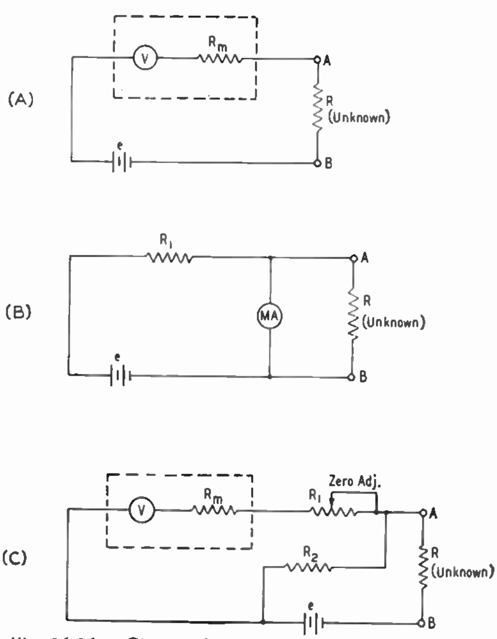


Fig. 16-16 — **Circuits for measuring resistance. Values are discussed in the text.**

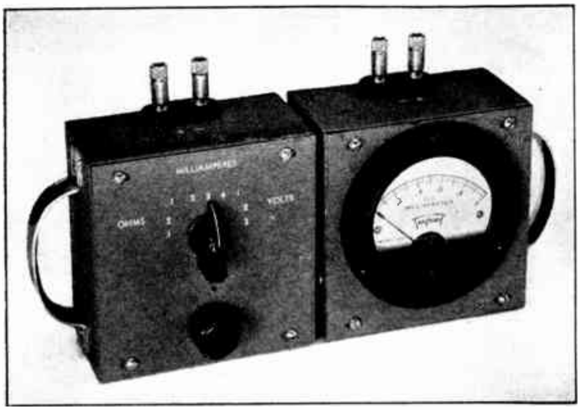


Fig. 16-17 — An inexpensive multirange volt-ohm-milliammeter. The 2 × 4 × 4-inch cabinet at the left houses the multipliers, shunts, switch and zero-adjustment resistor. The meter is mounted in the metal cabinet shown at the right. The units are provided with plugs and jacks so that the meter can be used independently or as the indicator component for other instruments. Connections to the volt-ohm-milliammeter, or to the meter alone, are made to the terminals mounted at the top of both boxes. Handles are mounted on the cabinets to facilitate handling.

Precision wire-wound resistors used as voltmeter multipliers cannot readily be made by the amateur because of the much higher resistance required (as high as several megohms). As an economical substitute, standard fixed resistors may be used. Such resistors are supplied in tolerances of 5, 10 or 20 per cent ± the marked values. By obtaining matched pairs from the dealer's stock, one of which is, for example, 4 per cent low while the other is 4 per cent high, and using the pairs in parallel or series to obtain the required value of resistance, good accuracy can be obtained at small cost. High-voltage multipliers are preferably made up of several resistors in series; this not only raises the breakdown voltage but tends to average out errors in the individual resistors attributable to manufacturing tolerances.

When d.c. voltage and current are known, the power in a d.c. circuit can be stated by simple application of Ohm's Law: $P = EI$. Thus the voltmeter and ammeter are also the instruments used in measuring d.c. power.

Multirange Voltmeters and Ohmmeters

A combination voltmeter-milliammeter having various ranges is extremely useful for experimental purposes and for trouble shooting in receivers and transmitters. As a voltmeter such an instrument should have high resistance so that very little current will be drawn in making voltage measurements. A voltmeter taking considerable current will give inaccurate readings when connected in a high-resistance circuit — for example, in various parts of a receiver. For such purposes the instrument should have a resistance of at least 1000 ohms per volt: a 0–1 milliammeter or 0–500 microammeter (0–0.5 ma.) is the basis of most multirange meters of this type. Microammeters having a range of 0–50 μa., giving a sensitivity of 20,000 ohms per volt, also are used.

The various current ranges on a multirange instrument can be obtained by using a number of shunts individually switched in parallel with the meter. A switch with low contact resistance must be used.

It is often necessary to check the value of a resistor or to find the value of an unknown resistance, particularly in receiver servicing. An **ohmmeter** is used for this purpose. The ohmmeter is a low-current d.c. voltmeter provided with a source of voltage (usually dry cells). In the simplest form, shown in Fig. 16-16A, the meter and battery are connected in series with the unknown resistance. If a given deflection is obtained with terminals *A-B* shorted, insertion of the resistance under measurement will cause the meter reading to decrease. When the resistance of the voltmeter is known, the following formula can be applied:

$$R = \frac{eR_m}{E} - R_m$$

where R is the resistance under measurement,
e is the voltage applied (A-B shorted),
E is the voltmeter reading with R connected, and
R_m is the resistance of the voltmeter.

The circuit of Fig. 16-16A is not suited to measuring low values of resistance (below a

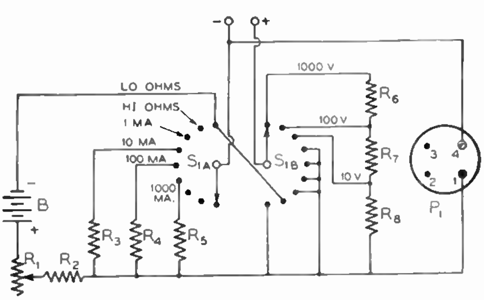


Fig. 16-18 — Diagram of the volt-ohm-milliammeter.

R_1 — 2000-ohm wire-wound variable.
R_2 — 3000 ohms, ½ watt.
R_3 — 10-ma. shunt, 6.11 ohms (see text).
R_4 — 100-ma. shunt, 0.555 ohm (see text).
R_5 — 1000-ma. shunt, 0.055 ohm (see text).
R_6 — 1000-volt multiplier, 0.9 megohm, ½ watt.
R_7 — 100-volt multiplier, 90,000 ohms, ½ watt.
R_8 — 10-volt multiplier, 10,000 ohms, ½ watt.
B — 4.5-volt dry battery (Burgess 5360).
P_1 — 4-prong male plug (for milliammeter).
S_{1A-B} — 9-point 2-pole selector switch (Mallory 3229J).

Fig. 16-19 — A rear view of the volt-ohm-milliammeter. The range-selector switch is mounted above the zero-adjustment potentiometer, and the shunts and multipliers are connected across the switch terminals. A four-prong male plug, for connection to the meter box, is shown at the left of the cabinet. The ohmmeter battery fits inside the case; the battery terminals should be insulated with tape or paper before the battery is installed in the box.

hundred ohms or so) with a high-resistance voltmeter. For such measurements the circuit of Fig. 16-16B can be used. The milliammeter should be a 0–1 ma. instrument, and R_1 should be equal to the battery voltage, e, multiplied by 1000. The unknown resistance is

$$R = \frac{I_2 R_m}{I_1 - I_2}$$

where R is the unknown.

R_m is the internal resistance of the milliammeter.

I_1 is the current in ma. with R disconnected from terminals A-B, and

I_2 is the current in ma. with R connected.

The formula is approximate, but the error will be negligible if e is at least 3 volts so that R is at least 3000 ohms.

A third circuit for measuring resistance is shown in Fig. 16-16C. In this case a high-resistance voltmeter is used to measure the voltage drop across a reference resistor, R_2, when the unknown resistor is connected so that current flows through it, R_2 and the battery in series. By suitable choice of R_2 (low values for low resistance, high values for high-resistance unknowns) this circuit will give equally good results on all resistance values in the range from one ohm to several megohms, provided that the voltmeter resistance, R_m, is always very high (50 times or more) compared with the resistance of R_2. A 20,000-ohm-per-volt instrument (50-μamp. movement) is generally used. Assuming that the current through the voltmeter is negligible compared with the current through R_2, the formula for the unknown is

$$R = \frac{eR_2}{E} - R_2$$

where R and R_2 are as shown in Fig. 16-16C.

e is the voltmeter reading with A-B shorted, and

E is the voltmeter reading with R connected.

The "zero adjuster," R_1, is used to set the voltmeter reading exactly to full scale when the meter is calibrated in ohms. A 10,000-ohm

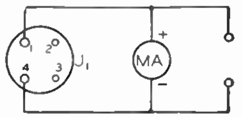


Fig. 16-20 — Wiring diagram of the 0–1 milliammeter shown in Figs. 16-9 and 16-17. J_1 is a 4-prong tube socket.

variable resistor is suitable with a 20,000-ohm-per-voltmeter. The battery voltage is usually 3 volts for ranges up to 100,000 ohms or so and 6 volts for higher ranges.

AN INEXPENSIVE V.O.M.

A combination multirange volt-ohm-milliammeter, reduced to simple and inexpensive terms, is shown in Figs. 16-17 to 16-20. Using a 0–1 milliammeter, the voltmeter has three

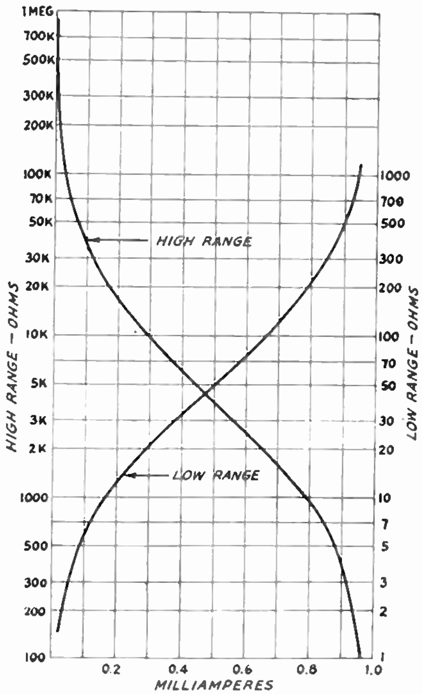


Fig. 16-21 — Calibration curve for the high- and low-resistance ranges of the volt-ohm-milliammeter.

ranges at 1000 ohms per volt: 0–10, 100 and 1000 volts. Current ranges of 0–1, 10, 100 and 1000 ma. are provided. There are two resistance-measurement ranges, a series range that is useful up to about 0.5 megohm, and a shunt range of 0–1000 ohms.

For economy, ordinary carbon resistors are used as voltmeter multipliers. These can be obtained with an accuracy within 5 per cent. However, standard resistors of 10 per cent tolerance can be used without introducing undue error. The 1000-volt multiplier, R_6, is two 1.8-megohm resistors connected in parallel, and the 100-volt multiplier, R_7, is two 0.18 megohm resistors arranged in parallel.

The 10-, 100- and 1000-ma. shunts are made of ordinary copper magnet wire wound on ½-watt resistors of high resistance value — 10,000 ohms or higher. The approximate lengths and sizes of the wire for the shunts are as follows: R_3, 9 feet No. 38 enameled; R_4, 5 feet No. 30 enameled; R_5, 8½ feet No. 18.

A calibration curve for the ohmmeter ranges is given in Fig. 16-21. With instruments having different internal resistance than the one shown in the photograph (Triplett Model 0321-1) the "low-ohms" curve will not apply exactly.

Grid-Dip Meters

A useful and inexpensive general-purpose instrument is an r.f. oscillator covering a wide frequency range. It generates signals that can be used for receiver alignment, for calibrating absorption wavemeters as described earlier in this chapter, and for furnishing small r.f. voltages for whatever purpose may be required. When equipped with a low-range milliammeter connected to read the oscillator grid current, it becomes a grid-dip meter and may be used for checking the resonant frequencies of tuned circuits, and as a means for measuring inductance and capacitance as described in a later section.

The grid-dip meter is so called because when its oscillator is coupled to a tuned circuit, the oscillator grid current will show a decrease or "dip" when the oscillator is tuned through resonance with the unknown circuit. The reason for this is that the external circuit will absorb energy from the oscillator when both it and the oscillator are tuned to the same frequency, and the loss of energy from the oscillator circuit causes the feed-back to decrease. The decrease in feed-back is accompanied by a decrease in grid current. The dip in grid current is quite sharp when the circuit to which the oscillator is coupled has reasonably high Q.

Any type of oscillator circuit can be used for the grip-dip meter, the only requirement being that a milliammeter of suitable range (0–1 is satisfactory in most cases) be connected in series with the grid leak. However, the grid-dip meter will be most useful when it covers a wide frequency range and is so constructed that it can be coupled to circuits in hard-to-reach places such as in a receiver chassis. The meters described in the following section have been designed with this in mind.

INEXPENSIVE GRID-DIP METER

The grid-dip meter shown in Fig. 16-22 is easy to build, handy to use, and covers a frequency range of 2850 kc. to 48 Mc. with five plug-in coils. This range readily can be extended in either direction, but for v.h.f. use a somewhat different version, shown later, is recommended. The circuit diagram of the oscillator is given in Fig. 16-23.

The support for the oscillator is a piece of aluminum measuring 9½ by 1½ inches, bent in the form of a "U" with sides 3¾ inches long so that the width of the "U" is just great enough (approximately 2 inches) for fastening to the mounting studs on the tuning condenser. As shown in Fig. 16-22, the socket for the plug-in coils is mounted across the open end of the "U" by means of small aluminum angle brackets. The socket for the 955 oscillator tube is similarly mounted near the closed end of the "U." The blocking and by-pass condensers are miniature ceramic units that take up very little space and thus contribute to compactness. The oscillator is provided with a handle (which can easily be made from a piece of broomstick) for ease of manipulation in checking circuits in receivers and transmitters.

The tuning condenser is a double-bearing unit originally of the single-section type having a maximum capacitance of 100 μμfd. To change

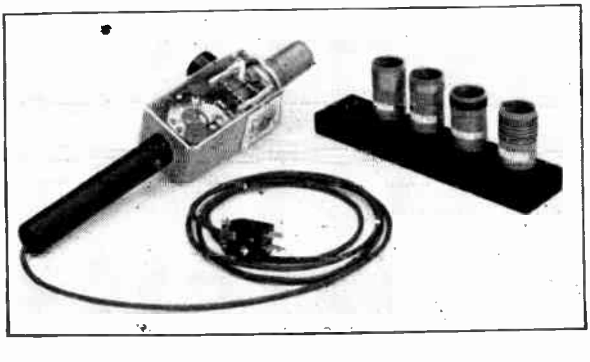


Fig. 16-22 — Inexpensive grid-dip oscillator using a 955 and plug-in coils. The five coils shown cover the range 2850 kc. to 48 Mc. An external 0–1 d.c. milliammeter is used as an indicator. Power and meter connections are brought through the four-wire cable.

it to the balanced type the center two stator plates are removed and the support bars sawed through at the middle. The rotor need not be touched. The stator plates can be removed without difficulty by bending them

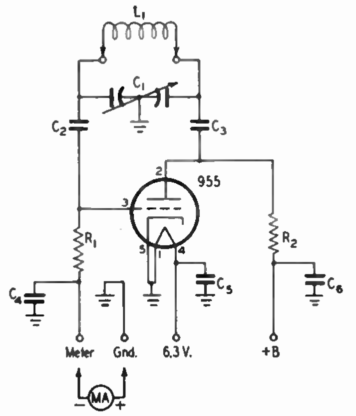


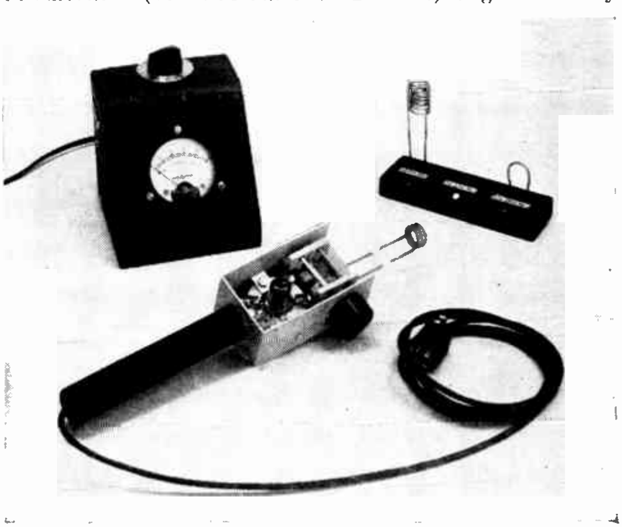
Fig. 16-23 — Circuit diagram of the grid-dip meter.
C_1 — Double-section midget, app. 12 $\mu\mu$fd. per section (Millen 21100 modified as described in text).
C_2, C_3 — 100-$\mu\mu$fd. ceramic (Centralab Hi-Kap).
C_4, C_5, C_6 — 0.01-μfd. ceramic (Sprague disc ceramic).
R_1 — 22,000 ohms, ½ watt, carbon.
R_2 — 68,000 ohms, ½ watt, carbon.
L_1 — 2.85–5.4 Mc.: 90 turns No. 30 s.c.c. on 1-inch form, close-wound.
— 4.6–8.7 Mc.: 37 turns No. 30 s.c.c. on 1-inch form, close-wound.
— 8.4–15.3 Mc.: 19 turns No. 30 s.c.c. on 1-inch form, close-wound.
— 14.4–25.5 Mc.: 11 turns No. 24 enam. on 1-inch form, close-wound.
— 25.1–48 Mc.: 8 turns No. 24 enam. on 1-inch form, spaced to occupy 13/16 inches.
MA — 0–1 d.c. milliammeter.

back and forth at the soldered joint with a pair of long-nose pliers until the solder breaks loose. The rotor should be grounded to the "U" frame at both ends; this helps to prevent dead spots (condenser settings at which the grid current shows rapid variations) in various portions of the range. The frequency calibration can be marked on a small piece of cardboard as shown in Fig. 16-22, using a pointer on the rear shaft extension of the condensers as an indicator.

♦

Fig. 16-24 — V.H.F. regenerative wavemeter/grid-dip meter, covering the 50–250 Mc. range. This is a high-sensitivity absorption-type wavemeter particularly useful for checking transmitter harmonics in television bands. The case in which the meter is mounted also contains the power supply. Regeneration is controlled by the knob on top of the case.

♦



The power requirements of the oscillator are 6.3 volts at 0.15 amp. for the 955 heater and a maximum of about 2 ma. at 150 volts for the plate. This power usually can be taken from a receiver or other existing supply. However, if a special supply is to be made for the instrument, the circuit of Fig. 16-27 will serve, the 1.5-volt dry cell shown in that diagram being omitted. In any event, it is a good idea to use a potentiometer, as shown in Fig. 16-27, for adjustment of plate voltage. In any grid-dip meter the grid current will be different in different parts of the frequency range, with fixed plate voltage, so that it is ordinarily necessary to choose a plate voltage that will keep the reading on scale in the part of the range where the grid current is highest. This usually results in rather low grid current at some other part of the range. With variable plate voltage this compromise is unnecessary.

The instrument may be calibrated by listening to its output with a calibrated receiver. High accuracy is not required in the applications for which a grid-dip meter is useful. The unit also may be used as an indicating wavemeter, in which case no plate voltage is needed since the grid and cathode of the 955 act as a simple diode. However, this type of circuit is not as sensitive as the crystal-detector type shown earlier in this chapter, because of the high-resistance grid leak in series with the meter.

REGENERATIVE WAVEMETER AND GRID-DIP METER

The unit shown in Fig. 16-24 is similar in construction to the grid-dip meter of Fig. 16-22, but in addition is an absorption wavemeter of very high sensitivity. The latter feature is particularly desirable in the v.h.f. range which this instrument covers, because of the necessity for detecting the presence of weak harmonics in the various television channels (54–88 Mc. and 174–216 Mc.). High sensitivity

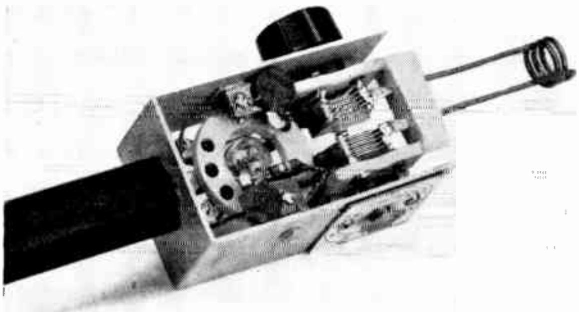


Fig. 16-25 — A bottom view of the regenerative wavemeter/grid-dip meter. This view shows the bottom of the 955 socket, with the miniature tubular ceramics mounted between the stator sections of the tuning condenser and the grid and plate terminals on the socket. The grid choke, shunting resistor, and by-pass condenser are at the bottom; the plate resistor, mounted through the socket, and the plate by-pass condenser are at the top. There is no wiring on the other side.

is achieved by operating the unit as a regenerative detector and by eliminating the grid-leak resistance, a low-resistance r.f. choke being substituted. The frequency range that can be covered satisfactorily with a given choke is limited, but the choke specified in the circuit diagram, Fig. 16-26, has been found to be adequate over the range 50–250 Mc.

With this instrument variable plate voltage is essential as a means of controlling regeneration. It is also essential to use the bias battery shown in the power-supply diagram of Fig. 16-27; without such bias there is a grid current of about 0.5 ma., even with no plate voltage on the tube, because of contact potential. Just as in the case of the lower-frequency instrument described earlier, the power for the oscillator can be taken from any existing supply. The plate-supply requirements are 150 volts and approximately 4 ma. About half of this current flows through the voltage divider, R_1, in Fig. 16-27.

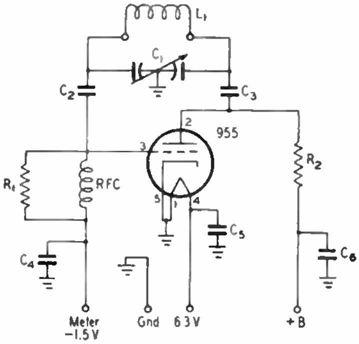


Fig. 16-26 — Circuit diagram of the regenerative wavemeter/grid-dip meter.

C₁ — Double-section midget, app. 36 μμfd. per section (Millen 21100 modified as described in text).
C₂, C₃ — 50-μμfd. ceramic (Centralab Hi-Kap).
C₄, C₅, C₆ — 0.001-μfd. ceramic (Sprague disc ceramic).
R₁ — 22,000 ohms, ½ watt, carbon.
R₂ — 68,000 ohms, ½ watt, carbon.
L₁ — 48–98 Mc.: 7¾ turns No. 12, ½-inch diam., 1 inch long, with 3½-inch leads.
— 76–156 Mc.: 2¾ turns No. 12, ½-inch diam., ⅜ inch long, 2½-inch leads.
— 130–265 Mc.: "U"-shaped loop, No. 12, 1½ inches long, ½ inch between sides.
RFC — Ohmite Z-144.

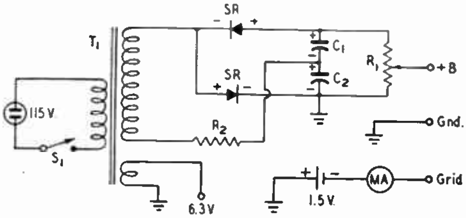


Fig. 16-27 — Power-supply circuit for the grid-dip meters shown in Figs. 16-22 and 16-24. When used with the meter of Fig. 16-22 the 1.5-volt battery should be omitted.

C₁, C₂ — 16-μfd. 150-volt electrolytic.
R₁ — 0.1-megohm potentiometer.
R₂ — 1000 ohms, 2 watts.
MA — 0–1 ma. (or smaller range for greater sensitivity).
S₁ — S.p.s.t. toggle (mounted on R_1).
SR — Selenium rectifier.
T₁ — Power transformer, required to furnish 6.3 volts at 0.3 amp. and app. 5 ma. at 115 volts (Millen 00011).

The tuning condenser, C_1, is the same type used in the instrument shown in Fig. 16-22 and is similarly modified into a split-stator unit. However, in this case a somewhat smaller minimum capacitance is desirable, so enough plates are removed from both rotor and stator so that each section consists of 5 stator and 5 rotor plates. Both ends of the rotor must be grounded to avoid dead spots. This can be done by soldering a short piece of wire between the contact washer and a mounting stud at each end. The ground connection is then made through the stud to the "U"-shaped support.

A crystal socket (half-inch spacing) with its lugs soldered directly to the condenser stators is used as a coil socket. No. 12 wire makes a good fit in such a socket, so the coils are self-supporting. A little additional strength for the socket mounting is secured by cementing it to the condenser end plates with Duco cement.

There are several methods by which the instrument can be given a frequency calibration. If a receiver is available covering at least a part of the range the unit can be used as an oscillator and calibrated against the receiver settings. Lecher wires also can be used; the method of using them is described earlier in this chapter.

To use the unit as a grid-dip meter the plate-voltage control is advanced to the point where a convenient value of grid current is obtained, after which it functions in the same way as the conventional grid-dip meter. To use it as a simple absorption wavemeter the plate voltage is turned off; the sensitivity under these conditions is about the same as the sensitivity of a crystal-detector wavemeter. To use it as a regenerative wavemeter the plate-voltage control is first advanced to the point where oscillation begins, as evidenced by a small amount of grid current, and then backed off until the grid current just disappears. This is the most sensitive condition. The setting of the plate-voltage control will depend to some extent on how tightly the instrument is coupled to the circuit being checked; tight coupling requires more plate voltage, loose coupling less. Care must be used to avoid false indications caused by actual oscillation should the coupling inadvertently be decreased; this usually can be checked by tuning over a small range about the desired frequency. When the unit is properly operated the grid current will show a sharp kick as the circuit is tuned through an actual signal and the current will drop to zero on either side. If the circuit is oscillating the grid current will be appreciable over a considerable tuning range.

Measuring Inductance and Capacitance

The ability to measure the inductance of coils, the capacitance of condensers, or the resonant frequency of a tuned circuit frequently saves time that might otherwise be spent in cut-and-try. A convenient instrument for this purpose is the grid-dip oscillator, described earlier in this chapter.

For measuring inductance, the coil is connected to a condenser of known capacitance as shown at A in Fig. 16-28. A mica condenser may be used as a standard; a 100-$\mu\mu$fd. 5 per cent tolerance unit will serve for most purposes. With the unknown coil connected to the standard condenser, the pick-up loop is coupled to the coil and the oscillator frequency adjusted for the grid-current dip, using the loosest coupling that gives a detectable indication. The inductance is then given by the formula

$$L_{\mu h.} = \frac{25{,}330}{C_{\mu\mu fd.}\, f_{Mc.}^{2}}$$

A calibrated variable condenser is generally used for measuring capacitance. The circuit is shown at B in Fig. 16-28. The frequency of the circuit, using any convenient coil, is first measured with the unknown capacitance disconnected and the calibrated condenser set near maximum. The unknown is then connected and the calibrated condenser readjusted to resonance. The unknown capacitance is then equal to the difference between the capacitances at the two settings of the calibrated condenser. Obviously only capacitances smaller than the maximum capacitance of the calibrated condenser can be measured by this method.

Since high accuracy in capacitance measurement is not ordinarily required, a satisfactory standard is any condenser of the straight-line capacitance type, for which a sufficiently good calibration curve can be constructed by noting the dial setting at which the plates just start to mesh and the setting at which they are completely meshed, and assuming that the capacitance change is linear within those limits. The minimum and maximum capacitance (corresponding closely enough to these condenser settings) can be obtained from the manufacturer's data on the particular variable condenser used.

An alternative method of measuring capacitance utilizes the fixed standard capacitance described above in inductance measurements, together with a coil of the proper inductance to resonate at a convenient part of the frequency range of the grid-dip meter. First measure the inductance of the coil with the standard condenser connected to it. Then substitute the unknown capacitance for the standard and determine the new resonant frequency. The unknown capacitance is then

$$C_{\mu\mu fd.} = \frac{25{,}330}{L_{\mu h}\, f_{Mc.}^{2}}$$

where f is the new frequency. This method is most adaptable to capacitances in the range 10–1000 $\mu\mu$fd. The standard condenser should be approximately 100 $\mu\mu$fd. for this range of measurement.

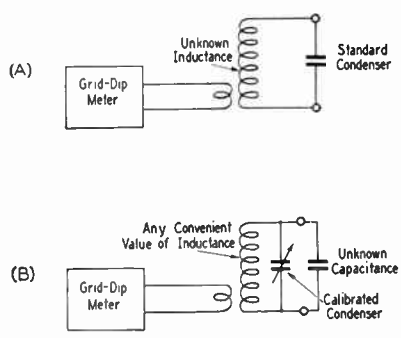


Fig. 16-28 — Set-ups for measuring inductance and capacitance with the grid-dip meter.

Audio-Frequency Oscillators

A useful accessory for testing audio-frequency amplifiers and modulators is an audio-frequency signal generator or oscillator. Checks for distortion, gain, and the ordinary troubles that occur in such amplifiers do not require elaborate equipment; in most cases, a single audio frequency in the 500–1000 cycle region will suffice. The chief requirement is that the audio oscillator be able to generate a reasonably good sine wave.

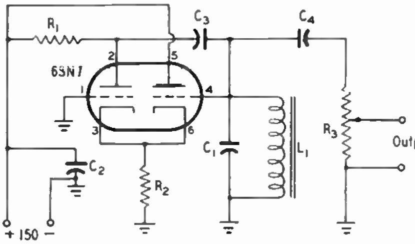


Fig. 16-29 — Audio oscillator circuit for fixed-frequency output.
C_1 — App. 0.05 μfd. (see text).
C_2 — 8-μfd. electrolytic.
C_3, C_4 — 0.1-μfd. paper.
R_1 — 68,000 ohms, 1 watt.
R_2 — 1500 ohms, 1 watt.
R_3 — 0.1-megohm potentiometer.
L_1 — App. 1 henry.

A circuit for a simple audio oscillator is given in Fig. 16-29. The second section of the 6SN7GT double triode is used to provide feed-back in the proper phase to the first section, so that oscillations can be maintained without requiring a tapped coil at L_1. The output amplitude is controlled by the potentiometer, R_3.

The frequency of oscillation is determined by L_1 and C_1. L_1 preferably should be an air-core coil, and can be an ordinary small "a.c.-d.c." filter choke with the iron core removed. Such coils usually will resonate in the vicinity of 400 cycles with 0.05-μfd. at C_1. If trial shows that the tone generated is too high or too low, appropriate changes in C_1 will bring it within the range desired. A number of frequencies can be made available by using several different values of capacitance, connected to a switch for convenient selection.

The output of such an oscillator with the control at maximum should be approximately 1.5 volts.

VARIABLE-FREQUENCY AUDIO OSCILLATOR

For measurements requiring a variable-frequency audio source the signal generator shown in Figs. 16-30 to 16-33, inclusive, is relatively inexpensive and easy to build. It uses a dual-triode oscillator with resistance-capacitance networks to obtain the phase shift required for oscillation, and includes an output amplifier and a power supply. It is built on a 7 × 9 × 2-inch chassis and housed in an 8 × 10 × 7-inch cabinet.

As indicated in the circuit diagram, Fig. 16-31, the oscillator tube is a 6SN7GT. The frequency of oscillation is determined by the resistance and capacitance in the network connected to the left-hand section of the oscillator tube. The small lamp, I_1, in the cathode lead to this tube section serves as a cathode bias resistor whose resistance varies with the oscillation amplitude in such a way as to maintain the output voltage essentially constant. The feed-back amplitude is controlled by R_{19}. With a variable condenser padded as specified in Fig. 16-31 the frequency range with a given set of resistors is somewhat less than 4 to 1, so five sets of resistors are needed to cover the 30–15,000 cycle range. S_1 selects the range required for a particular measurement.

The output of the oscillator is coupled through the amplitude control, R_{24}, to the 6J5 amplifier. Two output connections are provided. One, using the 6J5 as an ordinary amplifier, is for working into high-impedance circuits. The second, using the 6J5 as a cathode follower, is for circuits having an impedance in the neighborhood of a few thousand ohms. Shorting-type output jacks are used at J_1 and J_2 so that the unused output is properly by-passed by means of C_8 or C_7.

The power-supply section follows standard practice. Fairly good filtering is required, inasmuch as the oscillator covers the power-supply frequency range.

The construction of the instrument is shown in the photographs. The frequency-determining resistors, R_1 to R_{18}, are mounted on the terminals of the range switch. Placement of other parts is not critical. To reduce the possibility of trouble from a.c. hum, shielded wire is used for the heater circuits, the wiring to the

Fig. 16-30 — Variable-frequency audio signal generator. A dial of the type permitting direct calibration can be substituted if desired. This instrument is complete with power supply and covers the 30–15,000 cycle range in five steps.

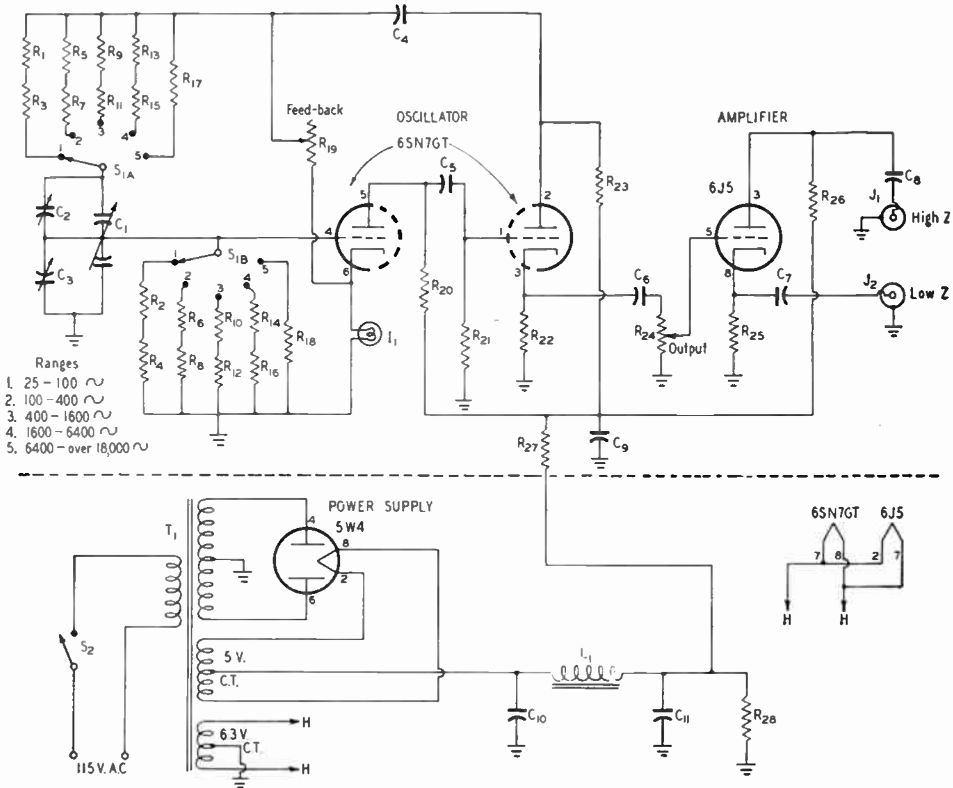


Fig. 16-31 — Circuit diagram of the audio-frequency signal generator.

C_1 — 450-μμfd.-per-section dual condenser, broadcast-receiver type.
C_2, C_3 — 130-μμfd. trimmer (100-μμfd. fixed mica in parallel with 3–30 compression trimmer).
C_4, C_8 — 20-μfd. electrolytic, 450 volts.
C_5 — 0.04-μfd. paper, 400 volts.
C_6, C_7 — 50-μfd. electrolytic, 25 volts.
C_9, C_{10} — 8-μfd. electrolytic, 450 volts.
C_{11} — 16-μfd. electrolytic, 450 volts.
R_1, R_2 — 8.2 megohms, 10% tolerance, ½ watt.
R_3, R_4 — 1.5 megohms, 10% tolerance, ½ watt.
R_5, R_6 — 2.2 megohms, 10% tolerance, ½ watt.
R_7, R_8 — 0.22 megohm, 10% tolerance, ½ watt.
R_9, R_{10} — 0.56 megohm, 10% tolerance, ½ watt.
R_{11}, R_{12} — 0.1 megohm, 10% tolerance, ½ watt.
R_{13}, R_{14} — 0.18 megohm, 10% tolerance, ½ watt.
R_{15}, R_{16} — 22,000 ohms, 10% tolerance, ½ watt.
R_{17}, R_{18} — 56,000 ohms, 10% tolerance, ½ watt.
R_{19} — 5000-ohm wire-wound potentiometer.
R_{20} — 47,000 ohms, 1 watt.
R_{21} — 1 megohm, ½ watt.
R_{22} — 1000 ohms, 1 watt.
R_{23} — 22,000 ohms, 1 watt.
R_{24} — 1-megohm potentiometer, audio taper.
R_{25} — 1500 ohms, 1 watt.
R_{26} — 56,000 ohms, 1 watt.
R_{27} — 10,000 ohms, 1 watt.
R_{28} — 60,000 ohms, 20 watts.
L_1 — 9 henrys, 50 ma. (Stancor C-1215).
I_1 — 4-watt 115-volt lamp.
J_1, J_2 — Shorting-type microphone jack (Amphenol 75-CL PC1M).
S_1 — 2-section 2-pole 5-position ceramic switch.
S_2 — S.p.s.t. switch (mounted on R_{24}).
T_1 — Power transformer, 650 volts c.t., 40 ma.; 5 volts, 3 amp.; 6.3 volts, 2 amp. (Stancor P-6010).

a.c. switch, and the leads to the output control, R_{24}. Good insulation is required in the frequency-determining RC circuits to avoid leakage, because of the high values of resistance required for the low-frequency ranges. A switch with ceramic wafers should be used at S_1, and the variable condenser should be mounted on ceramic button insulators.

The resistance values required for establishing adequate overlap between frequency ranges will not, in general, permit using single resistors of the preferred values. As shown in Fig. 16-31, two resistors are used in series in most cases so that, by combining preferred values appropriately, the desired resistance can be secured. Units with 10 per cent tolerance will be satisfactory. The exact value of total resistance is less important than that the corresponding pairs (e.g., R_1R_3 and R_2R_4) be matched as closely as possible to the same total resistance. Close matching is necessary to maintain the same oscillation amplitude on all frequency ranges. It is advisable to match the resistor pairs with an ohmmeter before installing them in the unit.

For preliminary alignment, set the trimmer condensers, C_2 and C_3, at maximum capacitance, set S_1 at Position 3, and connect a headset to the high-impedance output terminals. If R_9R_{11} and $R_{10}R_{12}$ are closely matched, an audio tone should be heard at some setting of R_{19}. Set R_{19} at the point that just maintains

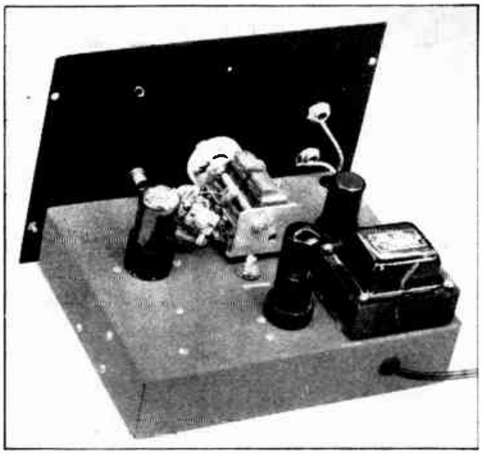


Fig. 16-32 — Rear chassis view of the audio signal generator. The oscillator tube is at the left, amplifier at the right. The feed-back control, screw-driver adjusted, is immediately in front of the tuning condenser in this view. The rectifier tube and power transformer are at the lower right.

oscillation, and turn C_1 through its entire range. If oscillations stop at any point adjust C_2 and C_3 for better capacity balance. When the proper trimmer settings have been obtained the oscillator will perform equally well at all settings of C_1.

Next, try the other frequency ranges and note which one gives oscillation at the lowest setting of R_{19}. If one or more ranges do not also work at this setting, the resistor pairs used on those ranges are not closely-enough matched. Other resistors of the same nominal rating should be substituted for the smaller values in each pair until oscillation is obtained on all ranges at the minimum setting of the feed-back control. This cut-and-try should not be required if the resistors are within 10 per cent tolerance. If a compromise setting of R_{19} has to be used, check the waveform on all ranges to make sure that it is satisfactory. The purest sine wave will be obtained when R_{19} is at the lowest setting that sustains oscillation. An oscilloscope will give the best check on waveform.

A fairly good frequency calibration can be secured by comparing the audio tone with the notes of the piano scale (see Chapter Twenty-Four). A more accurate calibration can be secured by comparing the signal with that from a good commercial unit. Alternatively, calibration points can be obtained with high accuracy throughout the entire range by using Lissajous patterns on an oscilloscope. The method is described in detail later in this chapter.

If R_{24} is set too high, the amplifier will overload and the output waveform will be distorted. At the highest setting that preserves the waveform, the open-circuit voltage at the high-impedance terminals is of the order of 60 volts, and at the low-impedance terminals is approximately 6 volts. A load of less than 1000 ohms on the low-impedance terminals will cause distortion. If the generator must work into a lower load impedance, the load can be connected in series with a resistor of the proper value to bring the total load up to at least 1000 ohms. Alternatively, a step-down transformer of the proper ratio may be connected between the generator and the load.

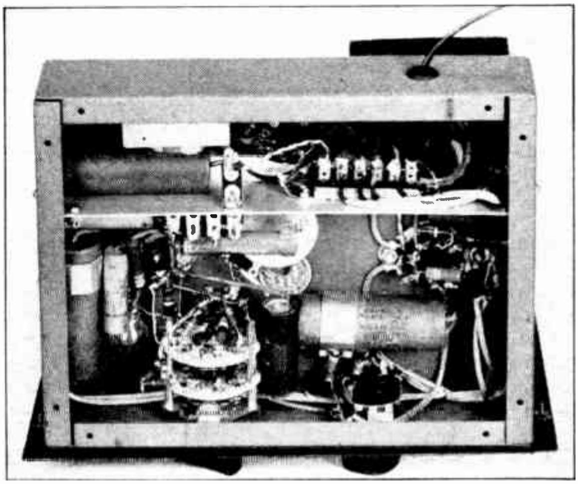


Fig. 16-33 — Below-chassis view of the audio signal generator. The regulator lamp is just to the left of the range switch, underneath the paper tubular, and is supported by terminals on the oscillator tube socket. The lengthwise strip of aluminum separates the power supply from the oscillator and amplifier, as a precaution against hum pick-up in the high-impedance RC circuits.

The Oscilloscope

The cathode-ray oscilloscope gives a visible representation of signals at both audio and radio frequencies and can therefore be used for many types of measurements that are not possible with instruments of the types described earlier in this chapter. For example, it can be made to show the waveform of an audio-frequency signal and thus detect distortion in an audio-frequency amplifier. With suitable calibration, it will measure a.c. voltages at radio as well as audio frequencies. The oscilloscope is such a versatile instrument that it is a highly valuable addition to the practical amateur station.

CATHODE-RAY TUBES

The heart of the oscilloscope is the **cathode-ray tube,** a vacuum tube in which the electrons emitted from a hot cathode are first accelerated to give them considerable velocity, then formed into a beam, and finally allowed to strike a special translucent screen which *fluoresces*, or gives off light at the point where the beam strikes. A narrow beam of moving electrons is analogous to a wire carrying current, and can be moved laterally, or **deflected,** by electric or magnetic fields.

Since the cathode-ray beam consists only of moving electrons, its weight and inertia are negligibly small. For this reason, it can be made to follow instantly the variations in periodically-changing fields at both audio and radio frequencies.

The electrode arrangement that forms the electrons into a beam is called the **electron gun.** In the simple tube structure shown in Fig. 16-34, the gun consists of the cathode, grid, and anodes Nos. 1 and 2. The intensity of the electron beam is regulated by the grid in the same way as in an ordinary tube. Anode No. 1 is operated at a positive potential with respect to the cathode, thus accelerating the electrons that pass through the grid, and is provided with small apertures through which the electron stream passes. On emerging from the apertures the electrons are traveling in practically parallel straight-line paths. The electrostatic fields set up by the potentials on anode No. 1 and anode No. 2 form an **electron lens** system which makes the electron paths converge to a point at the fluorescent screen. The potential on anode No. 2 is usually fixed, while that on anode No. 1 is varied to bring the beam into focus. Anode No. 1 is, therefore, called the **focusing electrode.**

Sharpest focus is obtained when the electrons of the beam have high velocity, so that relatively high d.c. potentials are common with cathode-ray tubes. However, the current required is small, so that the power consumption is negligible. A second grid may be placed between the control grid and anode No. 1, for additional acceleration of the electrons.

Methods of Deflection

When focused, the beam from the gun produces only a small spot on the screen, as described above. However, if after leaving the gun the beam is deflected by either magnetic or electrostatic fields, the spot will move across the screen in accordance with the force exerted on the beam. If the motion is rapid, the path of the spot **(trace)** appears as a continuous line.

Electrostatic deflection, the type generally used in the smaller tubes, is produced by **deflecting plates.** Two sets of plates are placed at right angles to each other, as indicated in Fig. 16-34. The fields are created by applying suitable voltages between the two plates of each pair. Usually one plate of each pair is connected to anode No. 2, to establish the polarities of the vertical and horizontal fields with respect to the beam and to each other.

Formation of Patterns

When periodically-varying voltages are applied to the two sets of deflecting plates, the path traced by the fluorescent spot forms a **pattern** that is stationary so long as the amplitude and phase relationships of the voltages remain unchanged. Fig. 16-35 shows how such patterns are formed. The horizontal sweep voltage is assumed to have the "sawtooth" waveshape indicated. With no voltage applied to the vertical plates the trace simply sweeps from left to right across the screen along the horizontal axis X–X' until the instant H is reached, when it reverses direction and returns

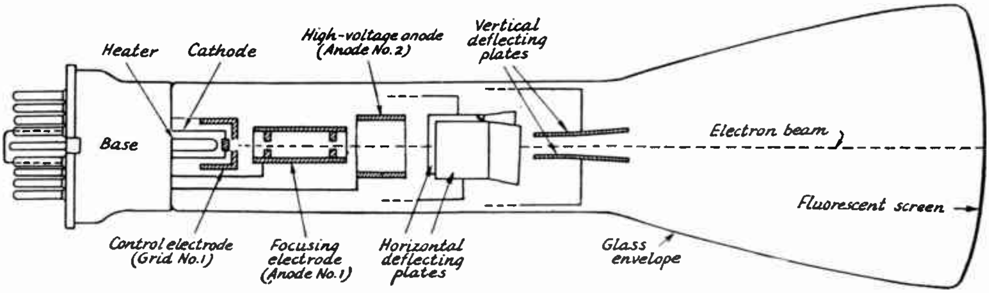


Fig. 16-34 — Typical construction for a cathode-ray tube of the electrostatic-deflection type.

to the starting point. The sine-wave voltage applied to the vertical plates similarly would trace a line along the axis $Y-Y'$ in the absence of any deflecting voltage on the horizontal plates. However, when both voltages are present the position of the spot at any instant depends upon the voltages on both sets of plates at that instant. Thus at time B the horizontal voltage has moved the spot a short distance to the right and the vertical voltage has similarly moved it upward, so that it reaches the actual position B' on the screen. The resulting trace is easily followed from the other indicated positions, which are taken at equal time intervals.

Types of Sweeps

A sawtooth sweep-voltage waveshape, such as is shown in Fig. 16-35, is called a linear **sweep,** because the deflection in the horizontal direction is directly proportional to time. If the sweep were perfect the **fly-back** time, or time taken for the spot to return from the end (H) to the beginning (I or A) of the horizontal trace, would be zero, so that the line HI would be perpendicular to the axis $Y-Y'$. Although the fly-back time cannot be made zero in practicable sweep-voltage generators it can be made quite small in comparison to the time of

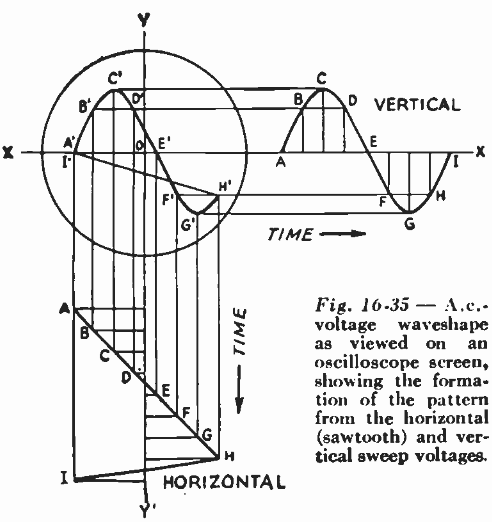


Fig. 16-35 — A.c.-voltage waveshape as viewed on an oscilloscope screen, showing the formation of the pattern from the horizontal (sawtooth) and vertical sweep voltages.

the desired trace AH, at least at most frequencies within the audio range. The fly-back time is somewhat exaggerated in Fig. 16-35, to show its effect on the pattern. The line $H'I'$ is called the **return trace**; with a linear sweep it is less brilliant than the pattern, because the spot is moving much more rapidly during the fly-back time than during the time of the main trace. If the fly-back time is short enough, the return trace will be invisible.

The linear sweep has the advantage that it shows the shape of the wave in the same way that it is usually represented graphically. If the time of one cycle of the a.c. voltage applied to the vertical plates is a fraction of the time taken to sweep horizontally across the screen, several cycles of the vertical or "signal" voltage will appear in the pattern. The shape of only the last cycle (or the last few cycles, depending upon the number in the pattern and the characteristics of the sweep) to appear will be affected by the fly-back in such a case.

The shape of the pattern obtained, with a given signal waveshape on the vertical plates, obviously will depend upon the shape of the horizontal sweep voltage. If the horizontal sweep is sinusoidal, the main and return sweeps each occupy the same time and the spot moves faster horizontally in the center of the pattern than it does at the ends. When two sinusoidal voltages of the same frequency are applied to both sets of plates, the pattern may be a straight line, an ellipse, or a circle, depending upon the amplitudes and phase relationships of the two voltages.

For many amateur purposes a satisfactory horizontal sweep is simply a 60-cycle voltage of adjustable amplitude. In modulation monitoring (described in Chapter Nine) audio-frequency voltage can be taken from the modulator to supply the horizontal sweep. For examination of audio-frequency waveforms, the linear sweep is essential. Its frequency should be adjustable over the entire range of audio frequencies to be inspected on the oscilloscope.

Lissajous Figures

When sinusoidal a.c. voltages are applied to the two sets of deflecting plates in the oscilloscope the resultant pattern depends on the relative amplitudes, frequencies and phase of the two voltages. If the relationship between these quantities is random the pattern is in continuous motion, but if the ratio between the two frequencies is constant and can be expressed in integers the pattern will be stationary. This makes it possible to use the oscilloscope for determining an unknown frequency, provided a variable frequency standard is available, or for determining calibration points for a variable-frequency oscillator if a few known frequencies are available for comparison.

The stationary patterns obtained in this way are called "Lissajous figures." Examples of some of the simpler Lissajous figures are given in Fig. 16-36. Patterns of the type shown in Fig. 16-36 are obtained when the two voltages have equal amplitudes; in case one has greater amplitude than the other the patterns will be elongated in the direction having the larger amplitude but will retain the same essential features. The form of the pattern for a fixed frequency ratio depends on the phase relationship between the two voltages; these figures are for a 90-degree phase difference.

In every case the patterns shown will be produced when the higher of the two frequencies is applied to the horizontal deflecting plates. Should the lower frequency be applied to the horizontal plates the pattern will be turned at right angles. The frequency ratio is found by counting the number of loops along two adjacent edges. Thus in the third figure from the top there are three loops along a horizontal edge and only one along the vertical, so the ratio of the horizontal frequency to the vertical frequency is 3 to 1. Similarly, in the fifth figure from the top there are four loops along the horizontal edge and three along the vertical edge, giving a ratio of 4 to 3. Assuming that the known frequency is applied to the horizontal plates, the unknown frequency is

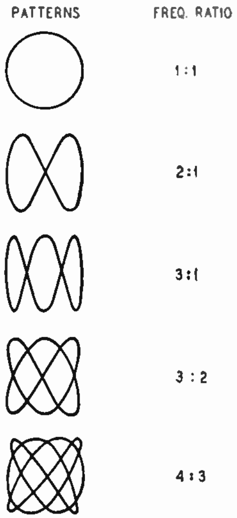


Fig. 16-36 — Lissajous figures and corresponding frequency ratios for a 90-degree phase relationship between the voltages applied to the two sets of deflecting plates.

$$f_2 = \frac{n_2}{n_1} f_1$$

where f_1 = known frequency applied to horizontal plates,
f_2 = unknown frequency applied to vertical plates,
n_1 = number of loops along a horizontal edge, and
n_2 = number of loops along a vertical edge.

In calibrating an oscillator, one of the frequencies is usually variable. The 90-degree pattern can be obtained by careful adjustment of the variable frequency until a stationary pattern resembling those shown is obtained. As the phase is varied the patterns will assume various forms, for a given frequency ratio, but the 90-degree pattern is easily identified because it is the most symmetrical.

An important application of Lissajous figures is in the calibration of audio-frequency signal generators, such as the variable-frequency a.f. oscillator described earlier in this chapter. Standard audio frequencies for this purpose are readily available. For very low frequencies the 60-cycle power-line frequency is held accurately enough to be used as a standard in most localities. The medium audio-frequency range can be covered by comparison with the 440-cycle modulation on the WWV transmissions, while high audio frequencies can be compared with WWV's 4000-cycle modulation. An oscilloscope having both horizontal and vertical amplifiers is desirable, since it is convenient to have a means for adjusting the voltages applied to the deflection plates to secure a suitable pattern size. The signal to the horizontal plates is fed directly to the amplifier, the horizontal linear sweep (if any) in the 'scope being switched out. The 60-cycle voltage can be obtained from the secondary of a filament transformer. The 440 and 4000 cycle voltages from the WWV signal can be taken from the headphone jack on a receiver. It is possible to calibrate over a 4-to-1 range, both upwards and downwards, from each of these three frequencies and thus cover the audio range completely.

A SIMPLE OSCILLOSCOPE FOR MODULATION MONITORING

Figs. 16-37 through 16-39 show the circuit and constructional details of a simple 2-inch oscilloscope that is suitable for use as a modulation monitor. It is designed to be mounted in the transmitter rack, becoming a permanent part of the 'phone station. Inexpensive parts are used throughout, and the circuits themselves are simple to build and operate.

The 2AP1 cathode-ray tube is mounted with its screen protruding through a 2-inch hole in the 19 × 5¼-inch aluminum rack panel. The cathode-ray tube is enclosed in a Millen shield, and its screen is covered by a Millen type 80072 bezel. The power-supply components are housed in a standard 3 × 4 × 5-inch utility box that is bolted to the left rear of the rack panel. An inexpensive replacement-type transformer is used with a 2X2 half-wave rectifier to deliver about 800 volts at the required 4 or 5 ma. drain.

The voltage-divider circuit components and the sweep-circuit controls are mounted on the right-hand side of the panel, and are enclosed by a 6 × 4¼ × 2½-inch three-sided box folded from sheet aluminum. A small audio transformer, mounted on the rear of this box, serves to provide 60-cycle sweep voltage. The by-pass condensers, C_2, C_3 and C_4, used to eliminate a.c. components from the d.c. control circuits, are connected directly to the rotor arms of their respective potentiometers, R_1, R_8 and R_9.

The socket for the cathode-ray tube is not fastened to any of the structural members of the unit but is used as a plug, with the socket

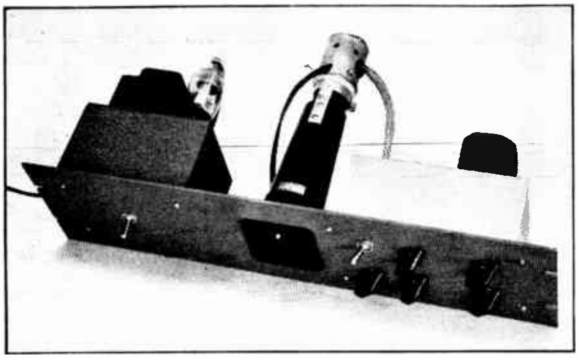


Fig. 16-37 — Front view of a rack-mounting oscilloscope for modulation monitoring. All components are mounted on the rear of a 19 × 5½-inch rack panel. The power-supply components are built into a utility box bolted on the left side of the panel, and the 'scope circuits are mounted on the right-hand side, enclosed by a shield box. The a.c. switch is on the left. All other controls are on the right, as follows: top row, l. to r., sweep switch, intensity control, focus control; bottom row, sweep-amplitude control, horizontal centering, vertical centering.

terminals enclosed in a tubular aluminum shield made by cutting down a National type T-78 tube shield. The base plate of this assembly is used as the support for a two-terminal tie point that holds isolating resistors R_{10} and R_{11}. These resistors are mounted inside the socket shield, as close to the tube base as possible. A ½-inch hole is drilled through the side of the shield to pass the cabled and shielded d.c. leads that run from the tube socket into the divider network in the aluminum shield box. A ceramic feed-through bushing requiring a ⅜-inch clearance hole passes through the opposite side of the socket shield to serve as the vertical input terminal. C_6 is connected between this bushing and the vertical deflection-plate pin on the tube socket. C_5, the coupling condenser for the horizontal plates, is mounted inside the larger shield compartment, near the horizontal-amplitude control, R_{12}.

The horizontal input terminals of the 'scope are mounted on the rear of the shield box, alongside of the audio transformer. The transformer secondary is connected to produce a turns ratio of approximately 1-to-1, which is sufficient to produce more than enough sweep voltage. A double-pole toggle switch is used to open the primary circuit of the audio transformer and to connect the external terminal to the amplitude control when the 'scope is used for transmitter monitoring. In this case sweep voltage is obtained from the audio system of the transmitter. R_{12} is connected on the tube side of the sweep switch, so that it remains in the circuit at all times to give control of voltage applied to the horizontal plates.

Details for using this oscilloscope to monitor a 'phone transmitter and to check both linearity and percentage modulation are contained in Chapter Nine. It should be remembered that an external resistor, R_E in Fig. 16-38, must be used in series with the lead to the horizontal

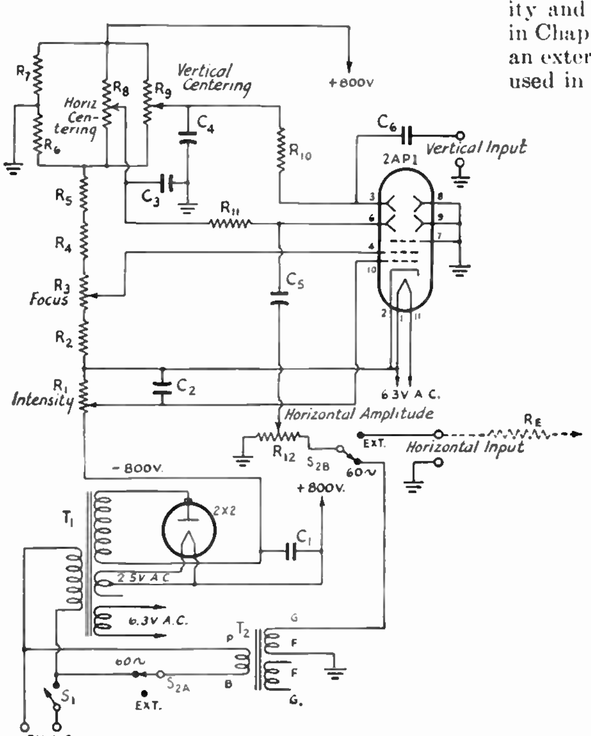


Fig. 16-38 — Circuit diagram of the simple oscilloscope for modulation monitoring.

C_1 — 1 μfd., 1000 volts, oil-filled.
C_2, C_3, C_4 — 0.01-μfd. 600-volt paper.
C_5 — 0.1 μfd., 1000 volts, paper.
C_6 — 0.001 μfd., 600 volts, mica.
R_1 — 20,000-ohm potentiometer, linear taper.
R_2 — 4700 ohms, ½ watt.
R_3 — 50,000-ohm potentiometer, linear taper.
R_4, R_5 — 33,000 ohms, 1 watt.
R_6, R_7 — 47,000 ohms, 1 watt.
R_8, R_9 — 50,000-ohm potentiometer, linear taper.
R_{10}, R_{11} — 1 megohm, ½ watt.
R_{12} — 0.25-megohm potentiometer, linear taper.
S_1 — S.p.s.t. toggle switch.
S_2 — D.p.d.t. toggle switch.
T_1 — Replacement-type receiver transformer, 350 v. each side of c.t., 70 ma. (Stancor P-6011.)
T_2 — Interstage audio transformer. (UTC S-2, with half of secondary unused; to produce approx. 1:1 turns ratio.)

Fig. 16-39 — Rear view of the rack-mounting oscilloscope. The shield covering the voltage-divider components has been removed to show wiring. Mounted on the shield are the audio transformer and the horizontal input terminals. The 'scope tube and its socket have been removed.

input terminals to reduce the audio voltage to the desired level. Instructions for selection of this resistor are given in Chapter Nine.

LINEAR SWEEPS AND AMPLIFIERS

Probably the chief use of the oscilloscope in amateur work is in measuring the percentage modulation in 'phone transmitters and in serving as a continuous monitor of modulation percentage. An oscilloscope for this purpose may be quite simple and inexpensive, consisting only of a small cathode-ray tube and an appropriate power supply as described earlier. However, by providing amplifiers for the deflection plates and furnishing a linear sweep circuit, the possibilities of the instrument are greatly extended. It then becomes possible, for example, to examine a.f. waveforms and to locate causes of distortion in a.f. amplifiers.

Gas-Tube Sweep Generator

A typical circuit for a linear sweep generator and amplifier is shown in Fig. 16-40. The tube is a gas triode or grid-control rectifier. The striking or breakdown voltage, which is the plate voltage at which the tube ionizes or "fires" and starts conducting, is determined by the grid bias. When plate voltage, E_b in Fig. 16-41, is applied, the condenser between plate and cathode acquires a charge through R_6R_7. The charging voltage rises relatively slowly, as shown by the solid line, until the breakdown or flashing point, V_f, is reached. Then the condenser discharges rapidly through the comparatively low plate-cathode resistance of the tube. When the voltage drops to a value too low to maintain plate-current flow, E_a, the ionization is extinguished and the condenser once more charges through R_6R_7. If the resistance is large enough, the voltage across

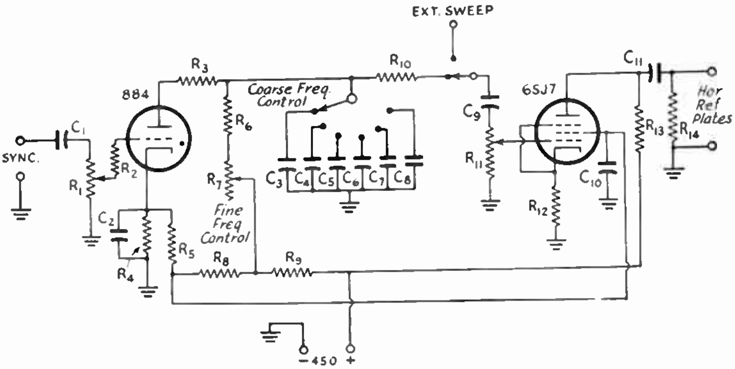


Fig. 16-40 — Linear sweep generator and horizontal amplifier.

C_1 — 0.1-μfd. paper.
C_2 — 25-μfd. 25-volt electrolytic.
C_3 — 0.25-μfd. paper, 600 volts.
C_4 — 0.1-μfd. paper, 600 volts.
C_5 — 0.01-μfd. paper, 600 volts.
C_6 — 0.015-μfd. paper, 600 volts.
C_7 — 0.005-μfd. paper or mica, 600 volts.
C_8 — 0.0022-μfd. mica.
C_9, C_{11} — 0.5-μfd. paper, 600 volts.
C_{10} — 8-μfd. electrolytic, 150 volts.
R_1 — 0.25-megohm potentiometer.
R_2 — 22,000 ohms, ½ watt.
R_3 — 470 ohms, ½ watt.
R_4 — 2200 ohms, ½ watt.
R_5 — 22,000 ohms, 1 watt.
R_6 — 0.33 megohm, ½ watt.
R_7 — 1-megohm potentiometer.
R_8, R_9 — 62,000 ohms, 1 watt.
R_{10} — 1 megohm, ½ watt.
R_{11} — 0.5-megohm potentiometer.
R_{12} — 820 ohms, ½ watt.
R_{13} — 0.1 megohm, 1 watt.
R_{14} — Bleed for horizontal deflection plates.

the condenser will rise linearly with time up to the breakdown point. This linear voltage change is used for the sweep. The fly-back time is the time required for condenser discharge through the sweep-generator tube; to keep this time small, the resistance during discharge must be low.

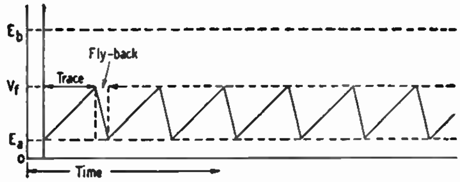


Fig. 16-41 — **Condenser charging curves showing how a sawtooth wave is produced by a gaseous-tube linear sweep oscillator.**

The "sawtooth" rate is controlled by varying the capacitance between plate and cathode and the resistance of R_6R_7. To obtain a stationary pattern, the sweep is synchronized by introducing some of the voltage being observed on the vertical plates into the grid circuit of the 884 gas triode. This voltage "triggers" the tube into operation in synchronism with the signal frequency. Synchronization will occur so long as the signal frequency is nearly the same as, or a multiple of, the self-generated sweep frequency.

The pentode amplifier in Fig. 16-40 can be used either to amplify the sweep-voltage output of the 884 oscillator, or to amplify any external voltage that it may be desired to use as a horizontal sweep. The gain control, R_{11}, provides a means for adjusting the width of the pattern on the cathode-ray tube screen. The output of the amplifier should be connected to the horizontal deflection plates of the tube. If this circuit is to be used with the oscilloscope previously described, the output terminals may be connected directly to Terminals 6 and 9 on the 2AP1 socket. In such case C_5 in Fig. 16-38 should be disconnected, but all other connections should be left unchanged.

Vertical Amplifiers

When using an oscilloscope for checking audio-frequency waveforms a "vertical" amplifier is a practical necessity. For most purposes the amplifier will be satisfactory if its frequency-response characteristic is flat over the a.f. range and if it has a gain of 100 or so. A typical circuit is shown in Fig. 16-42. It will be recognized as being practically similar to the "horizontal" amplifier of Fig. 16-40. A high-resistance gain control is desirable, to avoid loading the audio circuits to which the amplifier is connected.

When such an amplifier is used with the oscilloscope of Fig. 16-38, the output terminals should be connected between Terminals 3 and 8 on the 2AP1 socket. It is advisable to connect Terminal 3 to the arm of a 2-position ceramic switch, one contact going to the vertical amplifier and the other to C_6 in Fig. 16-38. This permits using either r.f. or a.f. input to the vertical deflection plates, disconnecting the a.f. amplifier circuit when r.f. voltage is to be applied.

Constructional Considerations

In building an oscilloscope, care should be taken to see that the tube is shielded from stray electric and magnetic fields that might deflect the beam, and means should be provided to protect the operator from accidental shock, since the voltages employed with the larger tubes are quite high. In general, the preferable form of construction is to enclose the instrument completely in a metal cabinet. From the standpoint of safety, it is good practice to provide an interlock switch that automatically disconnects the high-voltage supply when the cabinet is opened for servicing or other reasons.

In laying out the unit, the cathode-ray tube must be placed so that the alternating magnetic field from the power transformer has no effect on the electron beam. The transformer should be mounted directly behind the base of the tube, with the axes of the transformer windings and of the tube on a common line.

It is important that provision be included either for switching off the electron beam or reducing the spot intensity when no signal voltage is being applied. A thin, bright line or a spot of high intensity will "burn" the tube screen.

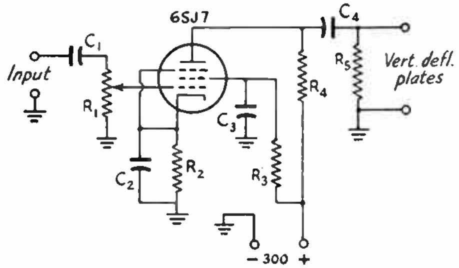


Fig. 16-42 — **Circuit diagram of a vertical amplifier for an oscilloscope.**

C_1, C_3, C_4 — 0.1-μfd. paper, 400 volts.
C_2 — 25-μfd. 25-volt electrolytic.
R_1 — 1-megohm potentiometer.
R_2 — 1500 ohms, ½ watt.
R_3 — 2.2 megohms, 1 watt.
R_4 — 0.47 megohm, 1 watt.
R_5 — Bleed resistor for vertical deflection plates.

If trouble is experienced in obtaining a clean pattern from a high-power transmitter because of r.f. voltage introduced by the 115-volt line, by-pass condensers (0.01 or 0.1 μfd.) should be connected in series across the primary of the power transformer, the common connection between the two being grounded to the oscilloscope case.

Antenna Measurements

Antenna measurements are made for the purpose (a) of securing maximum transfer of power to the antenna from the transmitter, and (b) of adjusting directional antennas to conform with design conditions. Measurements of the antenna system include the measurement of transmission-line performance.

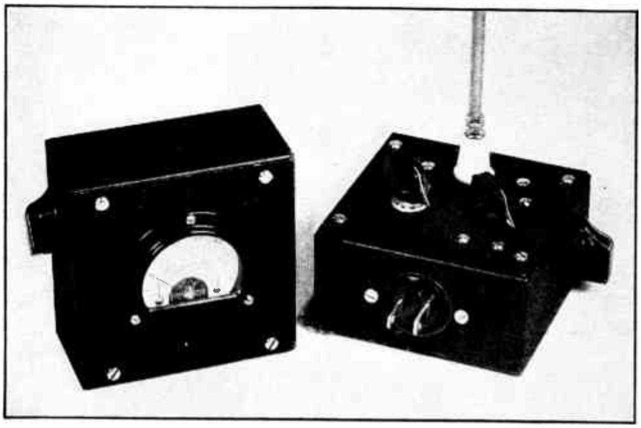


Fig. 16-43 — Remote-indicating field-strength meter, consisting of an r.f. pick-up and rectifier unit, and a meter unit. The knob on the left side of the meter unit is the switch for the shunt. On the pick-up unit the two controls are the bandswitch (left) and tuning. The knob at the right is for the resistor-shorting switch.

● FIELD-INTENSITY METERS

In adjusting antenna systems for maximum radiation and in determining radiation patterns, use is made of field-intensity meters. Fundamentally the field-intensity meter consists of a pick-up antenna and an indicating device such as a rectifier and microammeter, or a vacuum-tube voltmeter provided with a tuned input circuit. It is used to indicate the relative intensity of the radiation field under actual *radiating* conditions. It is particularly useful on the very-high frequencies and in adjusting directional antennas. Field-intensity checks should be made at points at least several wavelengths distant from the antenna and at heights corresponding with the desired angle of radiation.

The crystal-detector wavemeter described earlier in this chapter may be used as a field-strength meter if provided with a pick-up antenna. It is convenient to have the indicating device separate from the actual pick-up. This arrangement allows the pick-up unit to be set up out in the field to pick up radiation from the antenna under test, while the meter unit is near where adjustments are to be made. Antenna adjustment thus becomes a one-man job.

The unit shown in Figs. 16-43 to 16-45, inclusive, is particularly suitable for measurements in the v.h.f. range. It is constructed in two sections, one containing a tuned circuit, crystal rectifier, and antenna connection, and the other housing a microammeter for registering the rectified current from the crystal. The two units are fitted with matching plug and socket, permitting them to be used together, or they may be interconnected by means of a cable which can be any length up to several hundred feet. Three coils are used, so that measurements may be made on 28, 50 and 144 Mc. A resistor is inserted in series with the crystal and meter, to lessen the loading effect on the tuned circuit and to make the response of the crystal more linear with variations in received current. As the resistor reduces the sensitivity somewhat, a switch is provided to short it out in case measurements are to be made with extremely low power or at large distances from the transmitting antenna. A 100-microampere meter is used to give high sensitivity, and a shunt is available to multiply the range of the meter by three. This shunt is also provided with a switch so that low or high readings can be taken without making a trip to the pick-up unit. The crystal is the 1N21 type. Germanium crystals (1N34) also may be used with good results.

The two units are housed in 2 × 4 × 4-inch steel boxes with front and back removable.

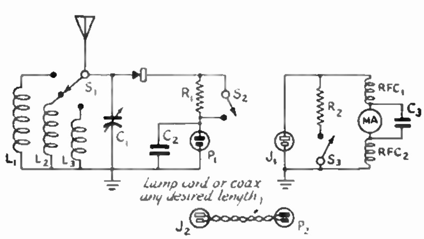


Fig. 16-44 — Wiring diagram of the remote-indicating field-strength meter.

C_1 — 25-μμfd. midget variable.
C_2, C_3 — 0.001-μfd. mica.
R_1 — 1000 ohms, ½ watt.
R_2 — 220 ohms, ½ watt.
L_1 — 28-Mc. coil — 7 turns No. 22 enamel, ¼ inch long, on ¾-inch dia. form (National PRF-1).
L_2 — 50-Mc. coil — 6 turns No. 22 enamel, ¼ inch long, on 9/16-inch dia. form (National PRE-1).
L_3 — 144-Mc. coil — 3 turns No. 18 enamel, ¼ inch long, ⅜-inch dia., self-supporting.
J_1, J_2 — Universal receptacle, two-pole retainer-ring type (Amphenol 61-F).
MA — 0–100 microammeter (0–500 microammeter or 0–1 milliammeter may be used, with reduced sensitivity).
P_1, P_2 — Polarized plug, two-pole retainer-ring type (Amphenol 61-MP).
S_1 — 3-position wafer-type switch.
S_2, S_3 — S.p.s.t. snap switch.
RFC_1, RFC_2 — 2.5 mh. choke (National R-100).

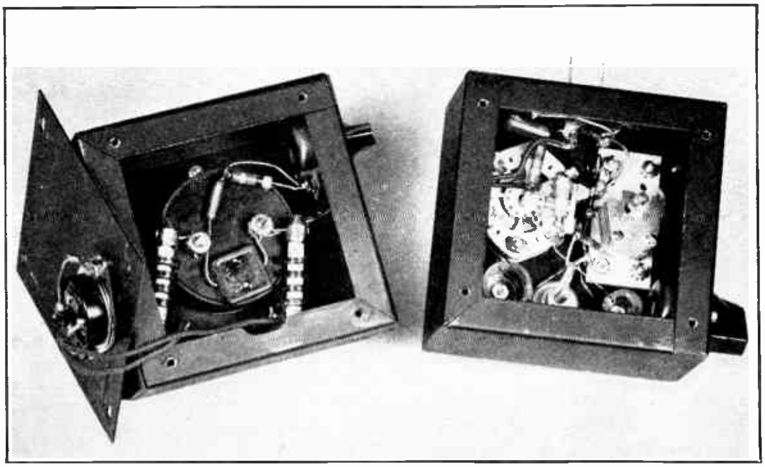


Fig. 16-45 — Inside view of the two units of the remote-indicating field-strength meter.

In the pick-up unit all parts except the resistor-shorting switch and connecting plug are mounted on the top panel, permitting easy wiring of the assembly. The interconnecting plug and socket are the polarized type, with one prong on the plug slightly larger than the other. The plug will fit a standard a.c. outlet, so the interconnecting cable (ordinary rubber-covered lamp cord) can double as a long a.c. extension cord.

The antenna connection is a steatite feed-through bushing fitted with a "banana-plug" socket. A convenient pick-up antenna is made by drilling and tapping a ¼-inch rod for 6/32 thread to take the threaded end of a banana plug. The length of the antenna will vary the sensitivity of the unit. If measurements are to be made with high power levels, a rod a few inches in length will suffice, but for ordinary work a 24-inch length will be suitable.

CHECKING STANDING WAVES

Standing waves on a transmission line can be measured if it is possible to measure the current at every point along the line, or the voltage between the two conductors at every point along the line. Rough checks on parallel-conductor lines can be made by going along the line with an absorption wavemeter having a crystal rectifier, taking care to keep the pick-up coil (or pick-up antenna) at the same distance from the line at every measurement. With such a device the milliammeter usually will indicate current loops if a small pick-up coil is used, and voltage loops if a short pick-up antenna is used.

An alternative indicator, also useful with parallel-conductor lines, is a neon lamp. With moderate amounts of transmitter power, a low-wattage lamp will glow when the glass bulb is brought into contact with one line wire. As the lamp is moved along the line, a change in brightness indicates standing waves. If the glow is substantially the same all along the line the s.w.r. can be considered to be low enough for practical purposes.

Standing-Wave Ratio Indicators

Simple indicators such as those just mentioned are useful for checking the presence of standing waves along a transmission line but are not adequate for actual measurement of the standing-wave ratio. In many cases, such

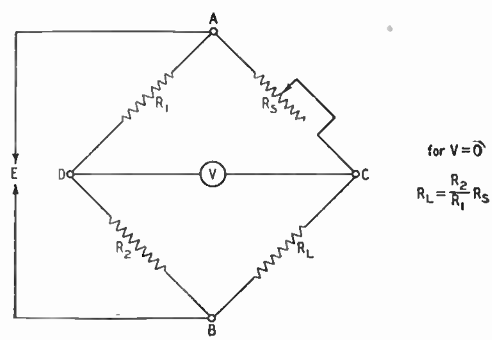


Fig. 16-46 — Resistance bridge as used for resistance measurement. This fundamental circuit is the basis for one type of bridge for measuring standing-wave ratio.

as adjustment of the match between a line and antenna, an accurate measurement is not at all necessary; it is sufficient simply to determine that a change in adjustment has either increased or decreased the s.w.r. But aside from accuracy, it is frequently inconvenient, and sometimes impossible, to move a current or voltage indicator along a transmission line for the distance required in checking standing waves.

An alternative method uses a bridge circuit to measure the standing-wave ratio, although not the standing waves themselves. While there are many forms of bridge circuits, the simple resistance bridge shown in Fig. 16-46 will serve to illustrate the basic principles. This type of bridge is often used for measurement of resistance. R_1 and R_2 are fixed resistors having known values, and R_S is a calibrated variable resistor. The unknown resistance to be measured, R_L, is connected in series with R_S to form a voltage divider across the source of voltage, E. The resistance of the voltmeter, V, should be very much larger than any of the four resistance "arms" of the bridge for maximum sensitivity. From Ohm's Law it is apparent that when R_1/R_2 equals R_S/R_L the voltage drops across R_1 and R_S are equal (this is also true of the voltage drops across R_2 and R_L) and there is no difference of potential between points C and D. Hence the voltmeter reading is zero ("null") and the bridge is said to be "balanced." Under any other conditions the potentials at C and D are not the same and the voltmeter reads the difference of potential. When the bridge is balanced,

$$R_L = R_S \frac{R_2}{R_1}.$$

R_1 and R_2 are called the "ratio arms" of the bridge. In bridges used for a.c. measurements two or more of the arms are frequently reactances (inductance or capacitance).

The basis for s.w.r. measurements with a bridge is the fact that the input impedance of a properly-terminated transmission line is a pure resistance equal to the line's characteristic impedance. If such a line is connected as the unknown arm of an appropriate bridge circuit the bridge can be balanced in the usual way and the indicating instrument will show a null. However, if the line is not properly terminated the input impedance will not equal the characteristic impedance, and will in general be reactive as well as resistive. Consequently, if the bridge is first balanced with a pure resistance equal to the characteristic impedance of the line, then on substituting the actual line the bridge will remain in balance only if the line is properly terminated — i.e., only if there are no standing waves. In all other cases the voltmeter will show an indication because the bridge has been thrown out of balance. It can be shown that this indication is a function of the standing-wave ratio, since the input impedance varies over a wider range as the s.w.r. increases. Hence the voltmeter can be calibrated in terms of s.w.r.

In addition to the resistance bridge shown in Fig. 16-46, the two circuits shown in Fig. 16-47 are well adapted to s.w.r. measurement. The one at A is a resistance-capacitance bridge and that at B a Maxwell-type bridge. All three bridges are theoretically independent of the applied frequency, and are practically so up to the frequency where skin effect, stray inductance, capacitance, and coupling between circuit elements and wiring become of importance. In both circuits the radio-frequency voltmeter, V, must be a high-impedance device. The conditions for "balance" — that is, for the voltmeter to read zero regardless of the voltage applied to the input terminals — are given in the equations to the right of each diagram. C_1 in Fig. 16-47A, and C in the circuit at B, are made adjustable so that the ratio of the bridge can be varied for various load resistances, R_L.

Practical circuits corresponding to the two in Fig. 16-47 are given in Figs. 16-48 and 16-49. The r.f. voltmeter is a crystal rectifier and 0–1 d.c. milliammeter (or microammeter) with chokes and resistors for keeping the r.f. out of the meter circuit. In order to keep the voltmeter impedance high and to improve the linearity, it is advisable to use as much resistance in series with the meter as possible while still obtaining full-scale indications at the r.f. power level used.

Several precautions must be observed in constructing and using such instruments. The leads must be kept short, to avoid introducing reactance that would prevent obtaining proper balance. The rectifier-circuit wiring should be kept out of the fields of the other components insofar as possible, since stray pick-up in this wiring will give a "residual" voltmeter reading that will not balance out. It is absolutely essential that the resistors have negligible capacitance and inductance: wire-wound resistors

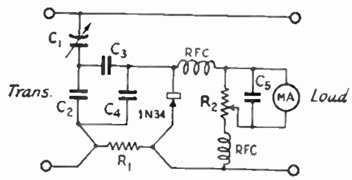


Fig. 16-48—Circuit diagram of the "Micro-Match" standing-wave indicator.

C_1 — 3-15-μμfd. midget variable.
C_2, C_4 — 220-μμfd. mica.
C_3 — 82-μμfd. mica.
C_5 — 0.0047-μfd. mica.
R_1 — 1.1-ohm resistor (nine 10-ohm 1-watt carbon resistors in parallel).
R_2 — 5000-ohm potentiometer.
MA — 0-1 d.c. milliammeter.
RFC — 2.5-mh. r.f. choke.

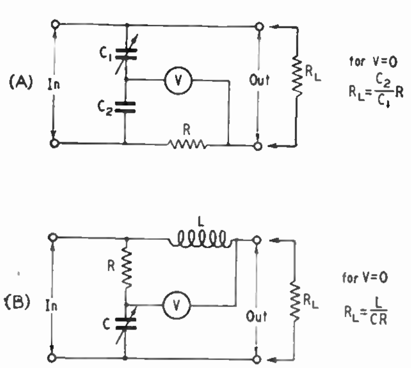


Fig. 16-47 — Fundamental circuits of two bridge-type standing-wave indicators. The upper circuit is used in the "Micro-Match" unit; the lower is a Maxwell bridge.

cannot be used with any degree of success.

To check either of the bridges shown in Figs. 16-48 and 16-49, connect a noninductive resistor equal to the characteristic impedance of the line to the output terminals, apply an r.f. voltage to the input terminals, and adjust the variable condenser for minimum reading.

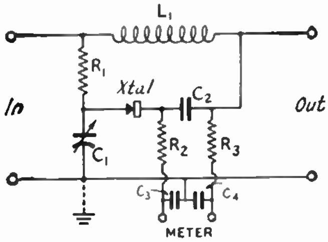


Fig. 16-49 — **Circuit diagram of the Maxwell-bridge standing-wave indicator. The meter should have a full-scale range of 1 milliampere or less.**

C_1 — 10–100-$\mu\mu$fd. Ceramicon variable.
C_2 — 470-$\mu\mu$fd. mica.
C_3, C_4 — (Optional) 100-$\mu\mu$fd. mica.
R_1 — 500 ohms, nonreactive.
R_2, R_3 — 10,000 ohms, ½-watt carbon.
L_1 — Approx. 29 turns No. 18, diameter 0.6 inch, 2.5 inches long.
XTAL — 1N34 or equivalent.

Then reverse the bridge so that the power source is connected to the output terminals and the resistor load to the input. Adjust the r.f. voltage (by changing the coupling to the transmitter) to make the meter read full scale. Then reverse the bridge connections and check the reading. If it is more than one or two per cent of the full-scale reading it will be necessary to try different arrangements of the wiring until the null reading can be brought as close to zero as possible.

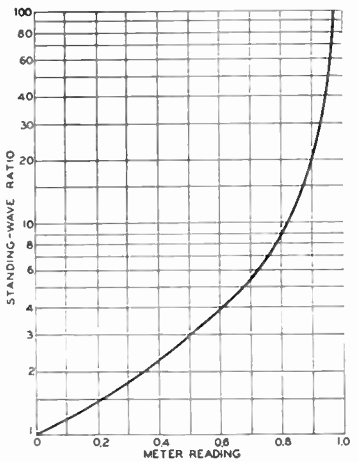


Fig. 16-50 — **Standing-wave ratio in terms of meter reading (relative to full scale) after setting outgoing voltage to full scale. This graph is a plot of the formula**

$$S.W.R. = \frac{V_o + V_r}{V_o - V_r}$$

where V_o and V_r are the outgoing and reflected components, respectively, of the voltage on the transmission line.

The variable condenser can be calibrated in terms of various line impedances by substituting load resistances of the appropriate values, noting the setting for balance at each resistance value. Both circuits can be used over the range of 50 to 300 ohms, approximately.

Calibration in terms of s.w.r. can be carried out, after checking for the null as described above, by using noninductive resistors of various values as loads. For a given line-impedance setting, the s.w.r. is given by

$$S.W.R. = \frac{R_L}{R_0} \text{ or } \frac{R_0}{R_L}$$

where R_0 is the line impedance for which the bridge has been adjusted to null, and R_L is the resistance used as a load. Use the formula that places the larger of the two resistances in the numerator. The theoretical calibration curve for a bridge is shown in Fig. 16-50, but this curve should be used only as a guide. It will not apply except in the case where the voltmeter impedance is infinite and the applied voltage remains constant regardless of how the bridge is connected.

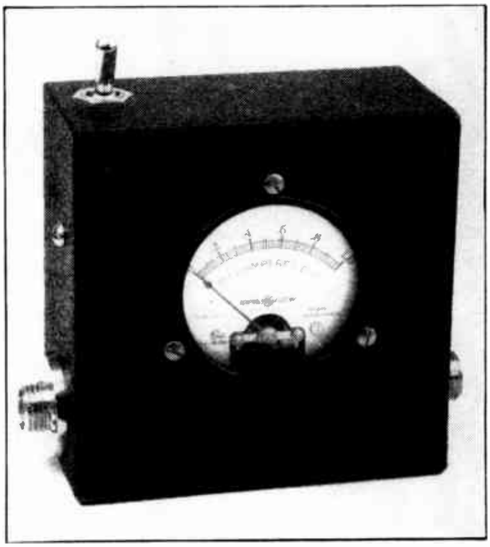


Fig. 16-51 — **Resistance-bridge standing-wave indicator for coaxial lines. Input and output terminals are at the lower left and lower right, respectively. This unit, built in a 2 by 4 by 4 box, is provided with a switch so that the voltmeter can measure either the applied voltage or the bridge voltage.**

To use either bridge for s.w.r. measurements after calibration, first reverse the bridge — that is, connect the line to the input terminals and the transmitter to the output terminals — and adjust the transmitter coupling to make the voltmeter read full scale. Then, leaving the transmitter coupling fixed, reconnect the bridge in the normal way, when the voltmeter will indicate the s.w.r.

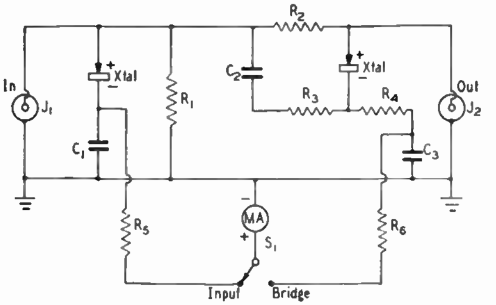


Fig. 16-52 — Resistance-bridge s.w.r. indicator for coaxial lines.

C_1, C_2, C_3 — 0.001-μfd. mica.
R_1 — App. 10 ohms, carbon, 5 watts (five 47-ohm 1-watt resistors in parallel).
R_2 — 50 to 75 ohms, carbon, ½ watt (select resistance value to equal characteristic impedance of line).
R_3, R_4 — App. 50 ohms. Absolute value not critical, but the two resistors should be within a few per cent of the same value.
R_5 — 4700 ohms, ½ watt, carbon.
R_6 — 820 ohms, ½ watt, carbon.
J_1, J_2 — Coax connectors.
MA — 0-1 d.c. milliammeter.
S_1 — S.p.d.t. toggle.
Xtal — 1N51 or 1N34.

Figs. 16-51 to 16-53, inclusive, show a resistance bridge built for coaxial lines, illustrating the type of construction that should be used in all types of bridge s.w.r. indicators. The construction should be such that coupling between various parts of the r.f. circuits is as small as possible. Short leads in the r.f. wiring are also important, to minimize stray reactances that, although not visible in the circuit diagram, may become appreciable at frequencies of the order of 14 Mc. and higher. In this bridge the two ratio arms (R_3 and R_4, Fig. 16-52) are equal, and this makes it unnecessary to reverse the bridge connections in either calibration or measurement. The loading resistor, R_1, is used principally to place a constant low-resistance load on the transmitter and thereby maintain constant voltage across the bridge regardless of the load that may be connected to the output terminals. An additional refinement, although not an essential part of the bridge, is the voltmeter connected across the input side of the line and consisting of the crystal rectifier, C_1, and R_5, in conjunction with S_1 and the milliammeter. This line voltmeter is a convenience in making measurements, because it will show whether or not the line voltage changes when shifting the output connections from open or short-circuit (the reference reading) to the actual line to be measured. Thus it shows whether or not an error has been introduced because of line voltage regulation, and permits readjustment to the proper value. The calibrations of the two voltmeters do not have to be identical.

The bridge performance can be checked by using a noninductive resistor of the same value (matched as closely as possible) as R_2 as a load. With the output terminals open and S_1 set to read input voltage, adjust the transmitter coupling to obtain a reading between half and full scale. Then connect the test resistor to the output terminals, using leads as short as possible, and readjust the transmitter coupling, if necessary, to maintain the same input voltage. Then switch S_1 to the bridge position, when the reading should drop to zero. A poor null under these conditions indicates stray coupling or excessive lead reactance in the bridge circuit.

The bridge may be calibrated by using noninductive resistors as described earlier. As a preliminary, adjust the transmitter coupling so that the voltmeter reads full scale (bridge position of S_1) with the output terminals open, and then check the input voltage. Connect various values of resistance across the output terminals, making sure that the input voltage is the same in each case, and note the reading with the meter in the bridge position. With this, as well as the other bridges, the readings may not correspond exactly for the same s.w.r. when appropriate resistors above and below the line impedance for which the bridge is designed are used. This is because of the current taken by the voltmeter. With the constants given in Fig. 16-52 the variation should not exceed about 5 per cent, and the

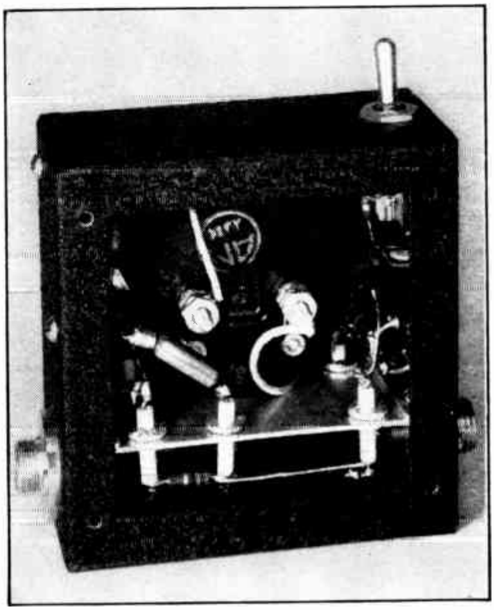


Fig. 16-53 — Inside view of the resistance-bridge s.w.r. indicator. The input terminal is at the right. An aluminum strip, the full width of the box, serves as a low-inductance ground plane for the instrument. Small ceramic through-bushings are used to insulate the "hot" line conductor and the bridge resistor, R_2 (Fig. 16-52), at the lower left. The ratio arms, R_3 and R_4, are mounted above the ground plane at the left edge of the box. The load resistor, at the right, consists of five 1-watt resistors mounted in ring fashion. This construction shields the hot conductor and bridge resistor from all other parts of the bridge.

error can be made smaller by using a low-range microammeter with a large series resistance as a voltmeter.

The procedure for using a resistance bridge for actual measurements is the same as that used during calibration.

S.w.r. measurements on parallel-conductor transmission lines are often subject to considerable error if there are appreciable "antenna" currents on the line (see Chapter Ten). These currents flow in parallel in the line conductors and affect the voltmeter independently of the true transmission-line currents. Their presence can be detected by interchanging the line wires at the output terminals. If the bridge does not give the same reading both ways the presence of antenna currents is indicated. In such a case neither reading is reliable. With coaxial lines antenna currents should give no trouble if the bridge is shielded and good solid connections are made at the input and output terminals.

The "Twin-Lamp"

A simple and inexpensive standing-wave indicator for 300-ohm line is shown in Fig. 16-54. It consists only of two flashlight lamps and a short piece of 300-ohm line. When laid flat against the line to be checked, the combination of inductive and capacitive coupling is such that outgoing power on the line causes the lamp nearest to the transmitter to light, while reflected power lights the lamp nearest the load. When the line is matched and no power is reflected, the lamp toward the antenna will be dark. The power input to the line should be adjusted to make the lamp nearest the transmitter light to full brilliance. When the lamp nearest the load just begins to glow, the s.w.r. is about 1.5 to 1.

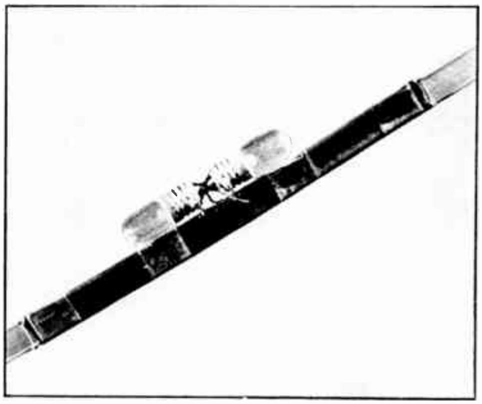


Fig. 16-54 — The "twin-lamp" standing-wave indicator.

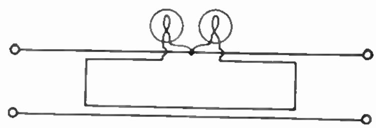


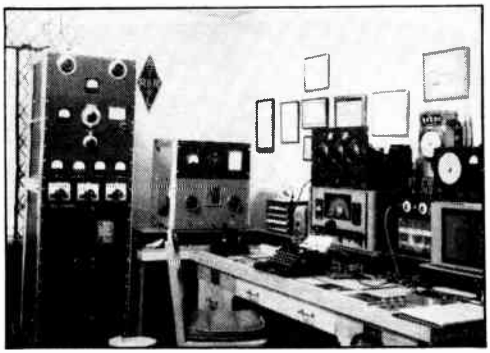
Fig. 16-55 — Wiring diagram of the "twin-lamp" standing-wave indicator.

To construct the "twin-lamp," take a short length (a foot or two) of 300-ohm Twin-Lead and remove about 1/4 inch of insulation from one wire at the center of the piece. Then take a second piece, 4 to 10 inches long (depending on the frequency and the transmitter power), and short-circuit both ends. Cut one wire in the exact center of the piece and peel the ends back on either side just far enough to provide leads to the flashlight lamps. Use the lowest-current flashlight bulbs or dial lamps available. Solder the tips of the bulbs together and connect them to the bare point in the long section of line, then solder the ends of the cut portion of the short piece to the shells of the bulbs. Figs. 16-54 and 16-55 should make the construction clear. The whole unit forms a "test section" that can be inserted in series with the line to be measured.

Assembling a Station

An amateur station is generally far better known by its signal and good operation than by its physical appearance. Good operating and a clean signal will build a reputation faster than thousands of dollars invested in special equipment and an elaborate "shack," and it is this very fact that makes amateur radio the democratic hobby that it is. However, most amateurs take pride in the arrangement of their stations, in the same way that they are careful of the appearance and arrangement of anything else which is part of the household. An antenna installation is the only external indication of the amateur station, and the degree of neatness required is generally determined by the district where the amateur lives and the attitude of the neighbors. However, with the advent of all different kinds of television receiving antennas, neighbors are in a much less favorable position to complain about the appearance of an amateur antenna system in the vicinity. TVI is something else, however!

The actual location inside the house of the "shack" — the room where the transmitter and receiver are located — depends, of course, on the free space available for amateur activities. Fortunate indeed is the amateur with a separate room that he can devote to his amateur station, or the few who can have a special small building separate from the main house. However, most amateurs must share a room with other domestic activities, and amateur stations will be found tucked away in a corner of the living room, a bedroom, a large closet, or even under the kitchen stove! A spot in the cellar or the attic can almost be classed as a separate room, although it may lack the "finish" of a normal room.



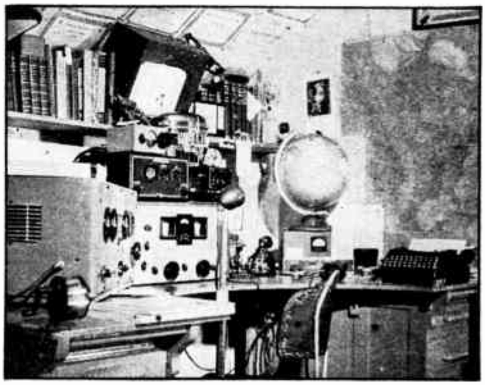
A good example of a station well prepared for activity on several bands. The rack houses power supply and 7- and 14-Mc. output amplifiers, with the 3.5-Mc. amplifier adjacent in its own rack. The receiver, VFO, tube keyer, typewriter, control switches, key and telephone are all within easy reach of the operator. Special cubbyholes provided for message forms, log book, Call Book and other papers keep the operating position neat and ready for action at any time. (W1CDA, Danville, Ky.)

Regardless of the location of the station, however, it should be designed for maximum operating convenience and safety. It is foolish to have the station arranged so that the throwing of several switches is required to go from "receive" to "transmit," just as it is silly to have the equipment arranged so that the operator is in an uncomfortable and cramped position during his operating hours. The reasons for building the station as safe as possible are obvious, if you are interested in spending a number of years with your hobby!

CONVENIENCE

The first consideration in any amateur station is the operating position, which includes the operator's table and chair and the pieces of equipment that are in constant use (the receiver, send-receive switch, and key or microphone). The table should be as large as possible, to allow sufficient room for the receiver or receivers, frequency-measuring equipment, monitoring equipment, control switches, and keys and microphones, with enough space left over for the logbook, a pad and pencil, and perhaps a *large* ash tray. Suitable space should be included for radiogram blanks and a call book, if these accessories are in frequent use. If the table is small, or the number of pieces of equipment is large, it is often necessary to build a shelf or rack for the auxiliary equipment, or to mount it in some less convenient location in or under the table. If one has the facilities, a semicircular "console" can be built of wood, or a simpler solution is to use two small wooden cabinets to support a table top of wood or Masonite. Home-built tables or consoles can be finished in any of the available oil stains, varnishes, paints or lacquers. Many operators use a large piece of plate glass over part of their table, since it furnishes a good writing surface and can cover miscellaneous charts and tables, prefix lists, operating aids, calendar, and similar accessories.

If the major interests never require frequent band changing, or frequency changing within a band, the transmitter can be located some distance from the operator, in a location where the meters can be observed from time to time (and the color of the tube plates noted!). If frequent band or frequency changes are a part of the usual operating procedure, the transmitter should be mounted close to the operator, either along one side or above the receiver, so that the controls are easily accessible without the need for leaving the operating position.



A convenient operating position can be obtained by building a "horseshoe-type" operating desk as shown here. Considerably more equipment can be placed on the desk around the operator than if an ordinary desk is used. (W9AND, Dixon, Ill.)

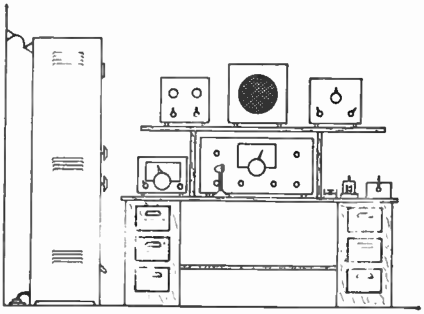


Fig. 17-1 — In a station assembled for maximum ease in frequency or band changing, the transmitter should be located next to the operating position, as shown above. On the operating table, the receiver is in front of the operator and VFO or crystal-switching oscillator on the left. (The VFO or crystal oscillator could be part of the transmitter proper, but most operators seem to prefer a separate VFO.)

The frequency standard and other auxiliary equipment can be mounted on a shelf above the receiver. The operating table can be an old desk, or a top supported by two small wooden cabinets. The "send-receive" switch is to the right of the telegraph keys — other switches are on the transmitter or the individual units.

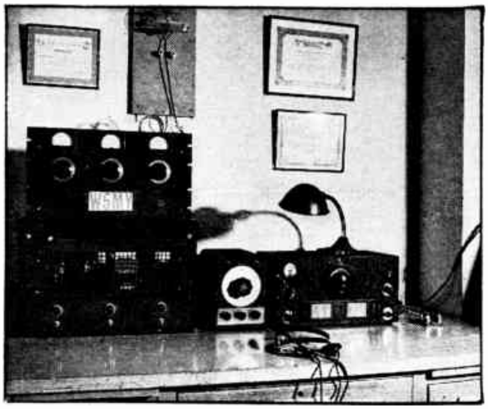
The above arrangement can be made to look cleaner by arranging all of the equipment on the table behind a single panel or a set of panels. In this case, provision must be made for getting behind the panel for servicing the units.

A compromise arrangement would place the VFO or crystal-switched oscillator at the operating position and the transmitter in some convenient location not adjacent to the operator. Since it is usually possible to operate over a portion of a band without retuning the transmitter stages, an operating position of this type is an advantage over one in which the operator must leave his position to make a change in frequency.

Controls

The operator has an excellent chance to exercise his ingenuity in the location of the operating controls. The most important controls in the station are the receiver tuning dial and the send-receive switch. The receiver tuning dial should be located four to eight inches above the operating table, and if this requires mounting the receiver off the table, a small shelf or bracket will do the trick. With the single exception of the amateur whose work is almost entirely in traffic or rag-chew nets, which require little or no attention to the receiver, it will be found that the operator's hand is on the receiver tuning dial most of the time. If the tuning knob is too high or too low, the hand gets cramped after an extended period of operating, hence the importance of a properly-located receiver. The majority of c.w. operators tune with the left hand, preferring to leave the right hand free for copying messages and handling the key, and so the receiver should be mounted where the knob can be reached by the left hand. 'Phone operators aren't tied down this way, and tune the communications receiver with the hand that is more convenient.

The hand key should be fastened securely to the table, in a line just outside the right



When one specializes in clean-cut c.w. operation on all bands, he is likely to come up with a neat arrangement like this. The transmitter runs 400 watts, despite its small size. The small unit between transmitter and receiver is the VFO. (W5MY, San Antonio, Texas.)

shoulder and far enough back from the front edge of the table so that the elbow can rest on the table. A good location for the semiautomatic or "bug" key is right next to the handkey, although some operators prefer to mount the automatic key in front of them on the left, so that the right forearm rests on the table parallel to the front edge.

The best location of the microphone is directly in front of the operator, so that he doesn't have to shout across the table into it, or run up the speech-amplifier gain so high that all manner of external sounds are picked up.

In any amateur station worthy of the name, it should be necessary to throw no more than one switch to go from the "receive" to the "transmit" condition. In 'phone stations, this switch should be located where it can be easily reached by the hand that isn't on the receiver. In the case of c.w. operation, this switch is most conveniently located to the right or left of the key, although some operators prefer to have it mounted on the left-hand side of the operating position and work it with the left hand while the right hand is on the key. Either location is satisfactory, of course, and the choice depends upon personal preference. Some operators use a foot-controlled switch, which is a convenience but doesn't allow too much freedom of position during long operating periods.

If the microphone is hand-held during 'phone operation, a "push-to-talk" switch on the microphone is convenient, but hand-held microphones tie up the use of one hand and

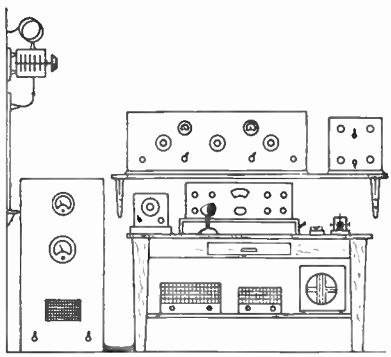


Fig. 17-2 — When little space is available for the amateur station, the equipment has to be spotted where it will fit. In the above arrangement, the transmitter, modulator and power supplies (separate units) are sandwiched in alongside the operating table and on a shelf above the table. The antenna tuning unit is mounted over the feed-through insulators that bring the antenna line into the "shack," and loudspeaker and small power supplies are mounted under the table. The operating position is clean, however, with the VFO, receiver and keys at table level. The tuning knob of this receiver would be uncomfortably low if the receiver weren't raised by the wooden arch, and the "send-receive" switch is mounted on the right-hand side of this arch, next to the hand key. Interconnecting leads should be cabled along the back of the table and table legs, to keep them inconspicuous.

This 700-watt station is tucked away in one corner of an apartment. The secret to a compact station is to do away with frills not necessary for communication. The transmitter and supply are in the cabinet, the VFO and receiver on the table. (W2ALO, Rutherford, N. J.)

are not too desirable, although they are widely used in mobile and portable work. A breast, chin or throat microphone is safer for mobile work, if the operator is also the driver of the vehicle.

The location of other switches, such as those used to control power supplies, filaments, 'phone/c.w. change-over and the like, is of no particular importance, and they can be located on the unit with which they are associated. This is not strictly true in the case of the 'phone/c.w. DX man, who sometimes has need to change in a hurry from c.w. to 'phone. In this case, the change-over switch should be at the operating table, although the actual change-over should be done by a relay that the switch controls.

If a rotary beam is used the control of the beam should be convenient to the operator. The beam-direction indicator, however, can be located anywhere within sight of the operator, and does not have to be located on the operating table unless it is small, or included with the beam control.

When several fixed beams are used, the selection of any one should be possible from the operating position, to minimize the time required to select the proper one. This generally means using a series of antenna relays or a stepping switch.

Frequency Spotting

In a station where a VFO is used, or where a number of crystals is available, the operator should be able to turn on only the oscillator of his transmitter, so that he can spot accurately his location in the band with respect to other stations. This allows him to see if he has anything like a clear channel (if such a thing exists in the amateur bands!), or to see what his frequency is with respect to another station. Such a provision can be part of the "send-receive" switch. Switches are available with a center "off" position, a "hold" position on one side,

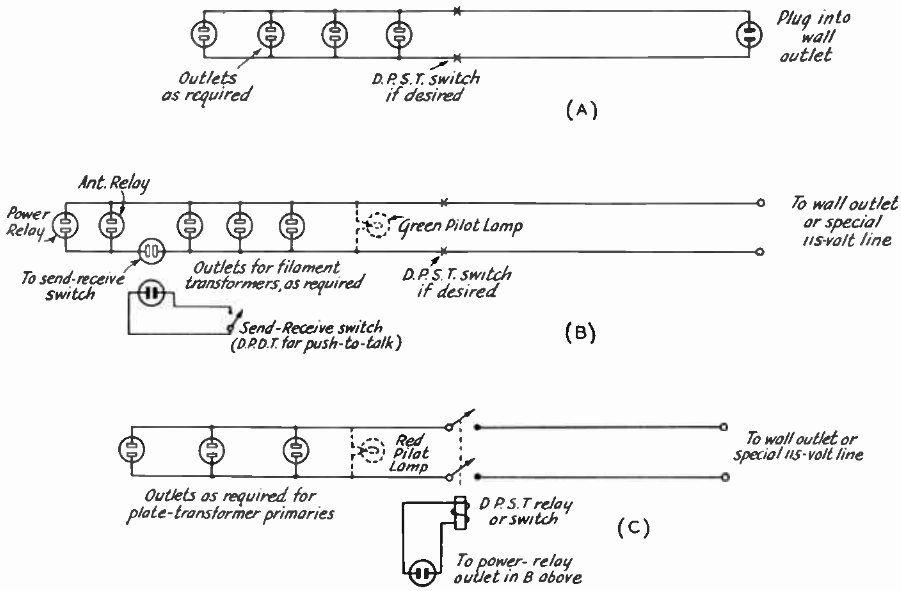


Fig. 17-3 — Power circuits for a high-power station. A shows the outlets for the receiver, monitoring equipment, speech amplifier and the like. The outlets should be mounted inconspicuously on the operating table. B shows the transmitter filament circuits and control-relay circuits, if the latter are used. C shows the plate-transformer primary circuits, controlled by the power relay. A heavy-duty switch can be used instead of the relay, in which case the antenna relay would be connected in circuit C.

If 115-volt pilot lamps are used, they can be connected as shown. Lower-voltage lamps must be connected across suitable windings on transformers.

With "push-to-talk" operation, the "send-receive" switch can be a d.p.d.t. affair, with the second pole controlling the "on-off" circuit of the receiver.

for turning on the oscillator only, and a "lock" position on the other side for turning on the transmitter and antenna relays. If oscillator keying is used, the key serves the same purpose, provided a "send-receive" switch is available to turn off the high-voltage supplies and prevent a signal going out on the air during adjustment of the oscillator frequency.

For 'phone operation, the telegraph key or an auxiliary switch can control the transmitter oscillator, and the "send-receive" switch can then be wired into the control system so as to control the oscillator as well as the other circuits.

An example of the compact station, complete on the operating table. The receiver is mounted on the left side of the table, for left-hand tuning. The beam-direction indicator and switches are housed in a small box sitting on the VFO. (W2NFU, Forest Hills, N. Y.)

Comfort

Of prime importance is the comfort of the operator. If you find yourself getting tired after a short period of operating, examine your station to find what causes the fatigue. It may be that the chair is too soft or hasn't a straight back or is the wrong height for you. The key or receiver may be located so that you assume an uncomfortable position while using them. If you get sleepy fast, the ventilation may be at fault. (Or you may need sleep!)

POWER CONNECTIONS AND CONTROL

Following a few simple rules in wiring your power supplies and control circuits will make it an easy job to change units in the station. If the station is planned in this way from the start, or if the rules are recalled when you are rebuilding, you will find it a simple matter to revise your station from time to time without a major rewiring job.

The regular wall outlets in a home are generally rated at 15 amperes at 115 volts, and so will furnish sufficient power for receivers, monitoring equipment, speech amplifiers, and anything that doesn't draw too high an intermittent load (such as a keyed transmitter or Class B modulator). A low-powered transmitter, under one or two hundred watts, can be supplied by an ordinary wall outlet. To make a neat installation, it is better to run a single pair of wires from the outlet over to the operating table or some central point, rather than to use a number of adapters at the wall outlet.

In a high-powered station, the receiver and auxiliary equipment can get their power from the wall outlet, but it is advisable to run in a special, heavy three-wire line from the meter box for the transmitter. This three-wire line will, of course, be 115 volts either side of neutral (ground), or 230 volts across the outside. In many cases it is possible to run the filaments and constant loads from one side of this three-wire line and the intermittent loads (plate transformers) from the other side. In this case the filament voltages will *rise* slightly with the application of load, because of the reduced *net* current in the neutral. However, this procedure often unbalances the system too much, resulting in considerable "blinking" of the lights, and the load must be distributed equally across the 230-volt circuit. This can be done by using plate transformers with 230-volt primaries, by dividing the load as equally as possible across both 115-volt circuits, or by using autotransformers that step down the 230 volts to 115 volts and connecting the plate-transformer primaries across the autotransformer secondaries. Obviously balancing the load is the cheapest "out" and the first one to try.

If the lights blink with keying or modulation of a low-powered transmitter that gets its power from a regular wall outlet, taking some of the power from another outlet may help to improve the regulation and is always worth a try.

When a special heavy line is run into the shack for a high-powered transmitter, it will generally be done by a licensed electrician who can advise you on the various types of outlets that are available. Some amateurs terminate their special lines in switch boxes, while others end the line in an electric-stove receptacle. In case you do the work yourself, it is wise to find out if there are any special regulations in your area covering the type of wire, insulation and outlet which must be used. The power companies are always willing to advise you if it looks as though you will be using more power!

Interconnections

The wiring of any station will entail two or three common circuits. The circuit for the receiver, monitoring equipment and the like, assuming it to be taken from a wall outlet, should be run from the wall to an inconspicuous point on the operating table, where it terminates in a multiple outlet large enough to handle the required number of plugs. A single switch between the wall outlet and the receptacle will then turn on all of this equipment at one time, or the plug can simply be pulled out at the wall when leaving the shack.

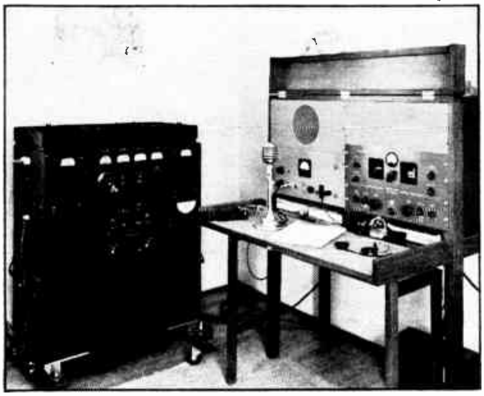
The second common circuit in the station is that supplying voltage to rectifier- and transmitter-tube filaments, bias supplies, and anything else that is not switched on and off during transmit and receive periods. The coil power for control relays should also be obtained from this circuit. The power for this circuit can come from a wall outlet or from the transmitter line, if a special one is used.

The third circuit is the one that furnishes power to the plate-supply transformers for the r.f. stages and for the modulator. When it is opened, the transmitter is disabled except for the filaments, and the transmitter should be safe to work on. However, one always feels safer when working on the transmitter if he has turned off every power supply pertaining to the transmitter.

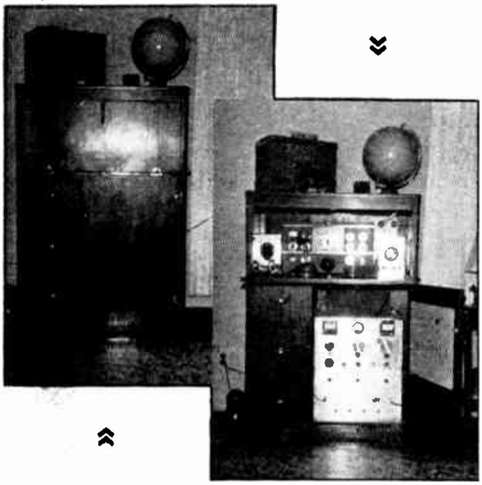
With these three circuits established, it becomes a simple matter to arrange the station for different conditions and with new units. Anything on the operating table (which runs all the time) ties into the first circuit. Any new power supply or r.f. unit gets its filament power from the second circuit. Since the third circuit is controlled by the send-receive switch (or relay), any power-supply primary that is to be switched on and off for send and receive connects to circuit No. 3.

Break-In and Push-To-Talk

In c.w. operation, "break-in" is any system that allows the transmitting operator to hear the other fellow's signal during the "key-up" periods between characters and letters. This allows the sending station to be "broken" by the receiving station at any time, to shorten calls, ask for "fills" in messages, and speed up operation in general. With present techniques, it requires the use of a separate receiving antenna and, with high power, some means for protecting the receiver from the transmitter when the key is "down." Several methods, applicable to high-power stations, are described in Chapter Eight. If the transmitter is low-powered (50 watts or so), no special



In this example of a compact high-power station, the operating table folds up when not in use and covers the receiver and speech amplifier. Special furniture, like this homemade operating table, goes a long way toward solving the space problem for many amateurs. (W4HAV, Fort Thomas, Ky.)



This station goes all the way in concealment by housing the entire station in a special cabinet. When the cabinet is opened, the operating table is formed and all pieces of gear are accessible. (W6YNX, Mountain View, Calif.)

equipment is required except the separate receiving antenna and a receiver that "recovers" fast. Where break-in operation is used, there should be a switch on the operating table to turn off the plate supplies when adjusting the oscillator to a new frequency, although during all break-in work this switch will be closed.

"Push-to-talk" is an expression derived from the "push" switch on some microphones, and it means a 'phone station with a single control for all change-over functions. Strictly speaking, it should apply only to a station where this single send-receive switch must be held in place during transmission periods, but any fast-acting switch will give practically the same effect. A control switch with a center "off" position, and one "hold" and one "lock" position, will give more flexibility than a straight "push" switch. The one switch must control the antenna change-over relay, the transmitter power supplies, and the receiver "on-off" circuit. This latter is necessary to disable the receiver during transmit periods, to avoid acoustic feed-back.

Switches and Relays

It is dangerous to use an overloaded switch in the power circuits. After it has been used for some time, it may fail, leaving the power on the circuit even after the switch is thrown to the "off" position. For this reason, large switches, or relays with adequate ratings, should be used to control the plate power. Relays are rated by coil voltages (for their control circuits) and by their contact ratings (the current they will carry safely).

When relays are used, the send-receive switch closes the circuit to their coils, thus closing the relay contacts. The relay contacts are in the power circuit being controlled, and thus the switch handles only the relay-coil current.

SAFETY

Of prime importance in the layout of the station is the personal safety of the operator and of visitors, invited or otherwise, during normal operating practice. If there are small children in the house, every step must be taken to prevent their accidental contact with power leads of any voltage. A locked room is a fine idea, if it is possible, otherwise housing the transmitter and power supplies in metal cabinets is an excellent, although expensive, solution. Lacking a metal cabinet, a wooden cabinet or a wooden framework covered with wire screen is the next-best solution. Many stations have the power supplies housed in metal cabinets in the operating room or in a closet or basement, and this cabinet or entry is kept locked — with the key out of reach of everyone but the operator. The power leads are run through conduit to the transmitter, using ignition cable for the high-voltage leads. If the power supplies and transmitter are in the same cabinet, a lock-type main switch for the incoming line power is a good precaution.

A simple substitute for a lock-type main switch is an ordinary line plug with a short connecting wire between the two pins. By wiring a female receptacle in series with the main power line in the transmitter, the shorting plug will act as the main safety lock. When the plug is removed and hidden, it will be impossible to energize the transmitter, and a stranger or child isn't likely to spot or suspect the open receptacle.

An essential adjunct to any station is a **shorting stick** for discharging any high voltage to ground before any work or coil changing is done in the transmitter. Even if interlocks and power-supply bleeders are used, the failure of one or more of these components may leave the transmitter in a dangerous condition. The shorting stick is made by mounting a small metal hook, of wire or rod, on one end of a dry stick or bakelite rod. A piece of ignition cable or other well-insulated wire is then run from the hook on the stick to the chassis or common ground of the transmitter, and the stick is hung alongside the transmitter. Whenever the power is turned off in the transmitter to work on the rig, or to change coils, the shorting stick is first used to touch the several high-voltage leads (tank condenser, filter condenser, tube plate connection, etc.) to insure that there is no high voltage at any of these points. Most commercial installations require the use of this simple device, and it has saved many a life. Use it!

Fusing

A minor hazard in the amateur station is the possibility of fire through the failure of a component. If the failure is complete and the component is large, the house fuses will generally blow. However, it is unwise and inconvenient to depend upon the house fuses to protect the lines running to the radio equipment, and every power supply should have its own set of fuses, with the fuse ratings selected at about 150 or 200 per cent of the maximum rating of the supply. If, for example, a power transformer is rated at 600 watts, it would draw about 5 amperes from the a.c. line ($600 \div 115 = 5.2$), and a 10-ampere fuse should be used in the primary circuit of the transformer. Circuit breakers can be used instead of fuses if desired.

Wiring

Control-circuit wires running between the operating position and a transmitter in another part of the room should be hidden, if possible. This can be done by running the wires under the floor or behind the base molding, bringing the wires out to terminal boxes or regular wall fixtures. Such construction, however, is generally only possible in elaborate installations, and the average amateur must content himself with trying to make the wires as inconspicuous as possible. If several pairs of leads must be run from the operating table to the transmitter, as is generally the case, a single piece of rubber- or vinyl-covered multiconductor cable will always look neater than several pieces of rubber-covered lamp cord.

The antenna wires always present a problem, unless coaxial-line feed is used. Open-wire line from the point of entry of the antenna line should always be arranged neatly, and it is generally best to support it at several points. Many operators prefer to mount their antenna-tuning assemblies right at the point of entry of the feedline, together with an antenna change-over relay (if one is used), and then the link from the tuning assembly to the transmitter can be made of inconspicuous coaxial line or Twin-Lead. If the transmitter is mounted near the point of entry of the antenna line, it simplifies the problem of "What to do with the feeders?"

There was enough room at this station to build the transmitter into the wall, and to protect it with glass doors. In an installation like this, it is convenient to have access to the rear of the transmitter units, for making connection to them and for testing. If the rear cannot be reached, all power leads should be cabled up along the side walls, at the rear. (W6NY, Whittier, Calif.)

General

You can check your station arrangement by asking yourself the following questions. If all of your answers are an honest "Yes," your station will be one of which you can be proud.

1) Is your station safe, under normal operating conditions, both for the operator and the visitor?

2) Is the operating position comfortable, even after several hours of operating?

3) Do you throw not more than one switch to go from "receive" to "transmit"?

4) Does it take only a short time to explain to another amateur how to work your station?

5) Do you show your station to visiting amateurs or laymen without apologizing for its appearance?

CHAPTER 18

The Amateur's Workshop

TOOLS AND MATERIALS

While an easier, and perhaps a better, job can be done with a greater variety of tools available, by taking a little thought and care it is possible to turn out a fine piece of equipment with only a few of the common hand tools. A list of tools which will be indispensable in the construction of radio equipment will be found on this page. With these tools it should be possible to perform any of the required operations in preparing panels and metal chassis for assembly and wiring. It is an excellent idea for the amateur who does constructional work to add to his supply of tools from time to time as finances permit.

Several of the pieces of light woodworking machinery, often sold in hardware stores and mail-order retail stores, are ideal for amateur radio work, especially the drill press, grinding head, band and circular saws, and joiner. Although not essential, they are desirable should you be in a position to acquire them.

INDISPENSABLE TOOLS

Long-nose pliers, 6-inch.
Diagonal cutting pliers, 6-inch.
Screwdriver, 6- to 7-inch, ¼-inch blade.
Screwdriver, 4- to 5-inch, ⅛-inch blade.
Scratch awl or scriber for marking lines.
Combination square, 12-inch, for laying out work.
Hand drill, ¼-inch chuck or larger, 2-speed type preferable.
Electric soldering iron, 100 watts.
Hack saw, 12-inch blades.
Center punch for marking hole centers.
Hammer, ball-peen, 1-lb. head.
Heavy knife.
Yardstick or other straightedge.
Carpenter's brace with adjustable hole cutter or socket-hole punches (see text).
Large, coarse, flat file.
Large round or rat-tail file, ½-inch diameter.
Three or four small and medium files—flat, round, half-round, triangular.
Drills, particularly ¼-inch and Nos. 18, 28, 33, 42 and 50.
Combination oil stone for sharpening tools.
Solder and soldering paste (noncorroding).
Medium-weight machine oil.

ADDITIONAL TOOLS

Bench vise, 4-inch jaws.
Tin shears, 10-inch, for cutting thin sheet metal.
Taper reamer, ½-inch, for enlarging small holes.
Taper reamer, 1-inch, for enlarging holes.
Countersink for brace.
Carpenter's plane, 8- to 12-inch, for woodworking.
Carpenter's saw, crosscut.
Motor-driven emery wheel for grinding.
Long-shank screwdriver with screw-holding clip for tight places.
Set of "Spintite" socket wrenches for hex nuts.
Set of small, flat, open-end wrenches for hex nuts.
Wood chisel, ½-inch.
Cold chisel, ½-inch.
Wing dividers, 8-inch, for scribing circles.
Set of machine-screw taps and dies.
Folding rule, 6-foot.
Dusting brush.

Twist Drills

Twist drills are made of either high-speed steel or carbon steel. The latter type is more common and will usually be supplied unless specific request is made for high-speed drills. The carbon drill will suffice for most ordinary equipment construction work and costs less than the high-speed type.

While twist drills are available in a number of sizes those listed in bold-faced type in Table 18-I will be most commonly used in construction of amateur equipment. It is usually desirable to purchase several of each of the commonly-used sizes rather than a quantity of odd sizes, most of which will be used infrequently, if at all.

Care of Tools

The proper care of tools is not alone a matter of pride to a good workman. He also realizes the energy which may be saved and the annoyance which may be avoided by the possession of a full kit of well-kept sharp-edged tools.

Drills should be sharpened at frequent intervals so that grinding is kept at a minimum each time. This makes it easier to maintain the rather critical surface angles required for best cutting with least wear. Occasional oilstoning of the cutting edges of a drill or reamer will extend the time between grindings.

The soldering iron can be kept in good condition by keeping the tip well tinned with solder and not allowing it to run at full voltage for long periods when it is not being used. After each period of use, the tip should be removed and cleaned of any scale which may have accumulated. An oxidized tip may be cleaned by dipping it in sal ammoniac while hot and then wiping it clean with a rag. If the tip becomes pitted, it should be filed until smooth and bright, and then tinned by dipping it in solder.

Useful Materials

Small stocks of various miscellaneous materials will be required in constructing radio apparatus, most of which are available from hardware or radio-supply stores. A representative list follows:

½ × 1/16-inch brass strip for brackets, etc. (half-hard for bending).
¼-inch-square brass rod or ½ × ½ × 1/16-inch angle brass for corner joints.
¼-inch diameter round brass rod for shaft extensions.
Machine screws: Round-head and flat-head, with nuts to fit. Most useful sizes: 4-36, 6-32 and 8-32, in lengths from ¼ inch to 1½ inches. (Nickel-plated iron will be found satisfactory except in strong r.f. fields, where brass should be used.)
Bakelite and hard-rubber scraps.
Soldering lugs, panel bearings, rubber grommets, terminal-lug wiring strips, varnished-cambric insulating tubing.

Machine screws, nuts, washers, soldering lugs, etc., are most reasonably purchased in quantities of a gross.

TABLE 18-I
Numbered Drill Sizes

Number	*Diameter (mils)*	*Will Clear Screw*	*Drilled for Tapping Iron, Steel or Brass**
1	228.0	—	—
2	221.0	12–24	—
3	213.0	—	14–24
4	209.0	12–20	—
5	205.0	—	—
6	204.0		—
7	201.0		—
8	199.0	—	—
9	196.0	—	—
10	193.5	10–32	—
11	191.0	10–24	—
12	189.0	—	—
13	185.0	—	—
14	182.0	—	—
15	180.0	—	—
16	177.0	—	12–24
17	173.0	—	—
18	**169.5**	**8–32**	—
19	166.0	—	12–20
20	161.0	—	—
21	159.0	—	10–32
22	157.0	—	—
23	154.0	—	—
24	152.0	—	—
25	149.5	—	10–24
26	147.0	—	—
27	144.0	—	—
28	**140.0**	**6–32**	—
29	**136.0**	—	**8–32**
30	128.5	—	—
31	120.0	—	—
32	116.0	—	—
33	**113.0**	**4–36, 4–40**	—
34	111.0	—	—
35	**110.0**	—	**6–32**
36	106.5	—	—
37	104.0	—	—
38	101.5	—	—
39	099.5	3–48	—
40	098.0	—	—
41	096.0	—	—
42	**093.5**	—	**4–36, 4–40**
43	089.0	2–56	—
44	086.0	—	—
45	082.0	—	3–48
46	081.0	—	—
47	078.5	—	—
48	076.0	—	—
49	073.0	—	2–56
50	**070.0**	—	—
51	067.0	—	—
52	063.5	—	—
53	059.5	—	—
54	055.0	—	—

*Use one size larger for tapping bakelite and hard rubber.

CHASSIS WORKING

With a few essential tools and proper procedure, it will be found that building radio gear on a metal chassis is no more of a chore than building with wood, and a more satisfactory job results.

The placing of components on the chassis is shown quite clearly in the photographs in this *Handbook*. Aside from certain essential dimensions, which usually are given in the text, exact duplication is not necessary.

Much trouble and energy can be saved by spending sufficient time in planning the job. When all details are worked out beforehand the actual construction is greatly simplified.

Cover the top of the chassis with a piece of wrapping paper or, preferably, cross-section paper, folding the edges down over the sides of the chassis and fastening with adhesive tape. Then assemble the parts to be mounted on top of the chassis and move them about until a satisfactory arrangement has been found, keeping in mind any parts which are to be mounted underneath, so that interferences in mounting may be avoided. Place condensers and other parts with shafts extending through the panel first, and arrange them so that the controls will form the desired pattern on the panel. Be sure to line up the shafts squarely with the chassis front. Locate any partition shields and panel

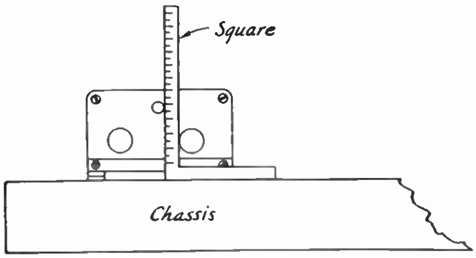


Fig. 18-1 — Method of measuring the heights of condenser shafts, etc. If the square is adjustable, the end of the scale should be set flush with the face of the head.

brackets next, and then the tube sockets and any other parts, marking the mounting-hole centers of each accurately on the paper. Watch out for condensers whose shafts are off center and do not line up with the mounting holes. Do not forget to mark the centers of socket holes and holes for leads under i.f. transformers, etc., as well as holes for wiring leads.

By means of the square, lines indicating accurately the centers of shafts should be extended to the front of the chassis and marked on the panel at the chassis line, the panel being fastened on temporarily. The hole centers may then be punched in the chassis with the center punch. After drilling, the parts which require mounting underneath may be located and the mounting holes drilled, making sure by trial that no interferences exist with parts mounted on top. Mounting holes along the front edge

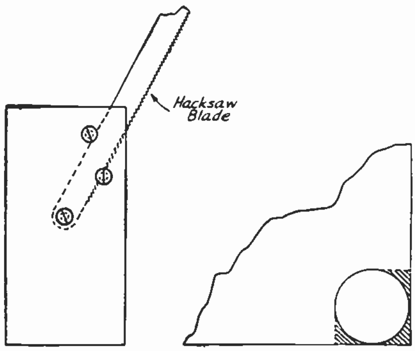


Fig. 18-2 — To cut rectangular holes in a chassis corner, holes may be filed out as shown in the shaded portion of B, making it possible to start the hack-saw blade along the cutting line. A shows how a single-ended handle may be constructed for a hack-saw blade.

of the chassis should be transferred to the panel, by once again fastening the panel to the chassis and marking it from the rear.

Next, mount on the chassis the condensers and any other parts with shafts extending to the panel, and measure accurately the height of the center of each shaft above the chassis, as illustrated in Fig. 18-1. The horizontal displacement of shafts having already been marked on the chassis line on the panel, the vertical displacement can be measured from this line. The shaft centers may now be marked on the back of the panel, and the holes drilled. Holes for any other panel equipment coming above the chassis line may then be marked and drilled, and the remainder of the apparatus mounted.

Drilling and Cutting Holes

When drilling holes in metal with a hand drill it is important that the centers first be located with a center punch, so that the drill point will not "walk" away from the center when starting the hole. When the drill starts to break through, special care must be used. Often it is an advantage to shift a two-speed drill to low gear at this point. Holes more than ¼ inch in diameter may be started with a smaller drill and reamed out with the larger drill.

The chuck on the usual type of hand drill is limited to ¼-inch drills. Although it is rather tedious, the ¼-inch hole may be filed out to larger diameters with round files. Another method possible with limited tools is to drill a series of small holes with the hand drill along the inside of the diameter of the large hole, placing the holes as close together as possible. The center may then be knocked out with a cold chisel and the edges smoothed up with a file. Taper reamers which fit into the carpenter's brace will make the job easier. A large rat-tail file clamped in the brace makes a very good reamer for holes up to the diameter of the file, if the file is revolved counterclockwise.

For socket holes and other large round holes, an adjustable cutter designed for the purpose may be used in the brace. Occasional application of machine oil in the cutting groove will help. The cutter first should be tried out on a block of wood, to make sure that it is set for the correct diameter. Probably the most convenient device for cutting socket holes is the socket-hole punch. The best type is that which works by turning a take-up screw with a wrench.

Rectangular Holes

Square or rectangular holes may be cut out by making a row of small holes as previously described, but is more easily done by drilling a ½-inch hole inside each corner, as illustrated in Fig. 18-2, and using these holes for starting and turning the hack saw. The socket-hole punch and the square punches which are now available also may be of considerable assistance in cutting out large rectangular openings. The burrs or rough edges which usually result after drilling or cutting holes may be removed with a file, or sometimes more conveniently with a sharp knife or chisel. It is a good idea to keep an old wood chisel sharpened and available for this purpose. A burr reamer will also be useful.

CONSTRUCTION NOTES

If a control shaft must be extended or insulated, a flexible shaft coupling with adequate insulation should be used. Satisfactory support for the shaft extension can be provided by means of a *metal* panel bearing made for the purpose. Never use panel bearings of the non-metal type unless the condenser shaft is grounded. *The metal bearing should be connected to the chassis with a wire or grounding strip.* This prevents any possible danger of shock.

The use of fiber washers between ceramic insulation and metal brackets, screws or nuts will prevent the ceramic parts from breaking.

Cutting and Bending Sheet Metal

If a sheet of metal is too large to be cut conveniently with a hack saw, it may be marked with scratches as deep as possible along the line of the cut on both sides of the sheet and then clamped in a vise and worked back and forth until the sheet breaks at the line. Do not carry the bending too far until the break begins to weaken; otherwise the edge of the sheet may become bent. A pair of iron bars or pieces of heavy angle stock, as long or longer than the width of the sheet, to hold it in the vise will make the job easier. "C"-clamps may be used to keep the bars from spreading at the ends. The rough edges may be smoothed up with a file or by placing a large piece of emery cloth or sandpaper on a flat surface and running the edge of the metal back and forth over the sheet.

Bends may be made similarly. The sheet should be scratched on both sides, but not so deeply as to cause it to break.

Cutting Threads

Brass rod may be threaded, or the damaged threads of a screw repaired, by the use of *dies*. Holes of suitable size (see Table 18-I) may be threaded for screws by means of *taps*. Taps and dies are obtainable in all standard machine-screw sizes. A set usually consists of taps and dies for 4–36 (or 4–40), 6–32, 8–32, 10–32 and 14–20 sizes, with a holder suitable for use with either tap or die. Machine oil applied to the tap usually makes cutting easier and sticking less troublesome.

Wiring

A popular type of wire for receivers is that known as "push-back" wire. It comes in sizes No. 16, 18, 20, etc., which are sufficiently large for all power circuits except filament. The insulating covering, which is sufficient for circuits where voltages do not exceed 400 or 500, can be pushed back a few inches at the end, making cutting of the insulation unnecessary when making a connection.

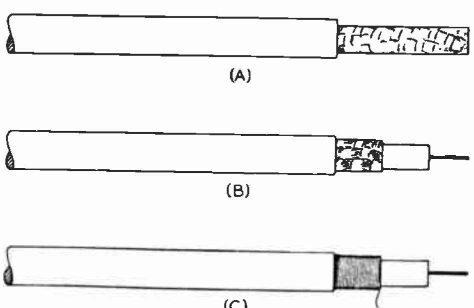


Fig. 18-3 — Method of preparing the end of coaxial cable. In A the outer insulating covering has been removed. At B the metal braid has been cut back, and in C the remaining exposed braid has been wrapped with small-size tinned wire. When completed, solder should be flowed over this winding.

Transmitter power wiring should be done with shielded wire, as discussed under "Harmonic Reduction," Chapter Six. Fig. 18-3 shows a common method of preparing the ends of shielded wire or cable. If the wire has an outer sheathing of insulation, this insulation should first be removed for a distance of about 2 inches, as shown at A. Then approximately the first inch of the shielding braid should be removed, as shown at B, by fraying the braid and cutting with diagonal cutters. At least a half inch of insulation should be left between the braid and the inner conductor. A solid connection can be made to the braid by winding a layer of No. 22 tinned wire over the braid, as shown at C, and then flowing solder over the winding. Care should be taken to prevent damage to the interior insulation during the soldering process. Filament wiring should be

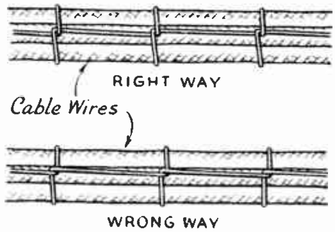


Fig. 18-4 — Right and wrong methods of lacing cable. With the right way the leading line is pinched under each turn and will not loosen if a break occurs in the lacing.

done with sufficiently large conductors to carry the required current without appreciable voltage drop (see Copper Wire Table, Chapter Twenty-Four). Rubber-covered house-wire sizes No. 14 to No. 10 are suitable for heavy-current transmitting tubes, while No. 18 to No. 14 flexible wire is satisfactory for receivers and low-drain transmitting tubes where the total length of the leads is not excessive.

Stiff bare wire, sometimes called **bus wire** or **bus bar**, is most favored for the high r.f. potential wiring of transmitters and, where practicable, in receivers. It comes in sizes No. 14 and No. 12 and is usually tin-dipped. Soft-drawn antenna wire also may be used. Kinks or bends can be removed by stretching 10 or 15 feet of the wire and then cutting it into small usable lengths.

The insulation of power wiring carrying high transmitter voltages should be appropriate for the voltage. Wire with rubber and varnished cambric covering, similar to ignition cable, is available from radio parts dealers.

The power-supply wiring should be done first. The leads should be bunched together as much as possible and kept down close to the surface of the chassis. The lacing of power wiring in cable form not only improves its appearance but also strengthens the wiring. Fig. 18-4 shows the correct procedure for lacing wires.

Chassis holes for wires should be lined with **rubber grommets** which fit the hole, to prevent chafing of the insulation. In cases where power-supply leads have several branches, it is often convenient to use **fiber terminal strips** as anchorages. These strips also form handy mountings for wire-terminal resistors, etc.

High-voltage wiring should have exposed points kept at a minimum and those which cannot be avoided rendered as inaccessible as possible to accidental contact.

Soldering

The secret of good soldering is in allowing time for the *joint*, as well as the solder, to attain sufficient temperature. Enough heat should be applied so that the solder will melt when it comes in contact with the wires being joined, without touching the solder to the iron.

Soldering paste, if of the noncorroding type, is extremely helpful when used correctly. In general, it should not be used for radio work except when necessary. The joint should first be warmed slightly and the soldering paste applied with a piece of wire. Only the bit of paste which melts from the warmth of the joint should be used. If the soldering iron is clean it will be possible with one hand to pick up a drop of solder on the tip of the iron which can be applied to the joint, while the other hand is used to hold the connecting wires together. The use of excessive soldering paste causes the paste to spread over the surface of adjacent insulation, causing leakage or breakdown of the insulation. Except where absolutely necessary, solder should never be depended upon for the mechanical strength of the joint; the wire should be wrapped around the terminals or clamped with soldering terminals.

TABLE 18-II

Standard Component Values

20% Tolerance	*10% Tolerance*	*5% Tolerance*
10	10	10
		11
	12	12
		13
15	15	15
		16
	18	18
		20
22	22	22
		24
	27	27
		30
33	33	33
		36
	39	39
		43
47	47	47
		51
	56	56
		62
68	68	68
		75
	82	82
		91
100	100	100

COMPONENT VALUES

Values of composition resistors and small condensers (mica and ceramic) are specified throughout this *Handbook* in terms of "preferred values." In the preferred-number system, all values represent (approximately) a constant-percentage increase over the next lower value. The base of the system is the number 10. Only two significant figures are used. Table 18-II shows the preferred values based on tolerance steps of 20, 10 and 5 per cent. All other values are expressed by multiplying or dividing the base figures given in the table by the appropriate power of 10. (For example, resistor values of 33,000 ohms, 6800 ohms, and 150 ohms are obtained by multiplying the base figures by 1000, 100, and 10, respectively.)

"Tolerance" means that a variation of plus or minus the percentage given is considered satisfactory. For example, the actual resistance of a "4700-ohm" 20-per-cent resistor can lie anywhere between 3700 and 5600 ohms, approximately. The permissible variation in the same resistance value with 5-per-cent tolerance would be in the range from 4500 to 4900 ohms, approximately.

Only those values shown in the first column of Table 18-II are available in 20-per-cent tolerance. Additional values, as shown in the second column, are available in 10-per-cent tolerance; still more values can be obtained in 5-per-cent tolerance.

In the component specifications in this *Handbook*, it is to be understood that when no tolerance is specified the *largest* tolerance available in that value will be satisfactory.

Values that do not fit into the preferred-number system (such as 500, 25,000, etc.) easily can be substituted. It is obvious, for example, that a 5000-ohm resistor falls well within the tolerance range of the 4700-ohm 20-per-cent resistor used in the example above. It would not, however, be usable if the tolerance were specified as 5 per cent.

COLOR CODES

Standardized color codes are used to mark values on small components such as composition resistors and mica condensers, and to identify leads from transformers, etc. The resistor-condenser number color code is given in Table 18-III.

Fixed Condensers

The methods of marking "postage-stamp" mica condensers, molded paper condensers, and tubular ceramic condensers are shown in Fig. 18-5. Condensers made to American War Standards or Joint Army-Navy specifications are marked with the 6-dot code shown at the top. Practically all surplus condensers are in this category. The 3-dot RMA code is used for condensers having a rating of 500 volts and

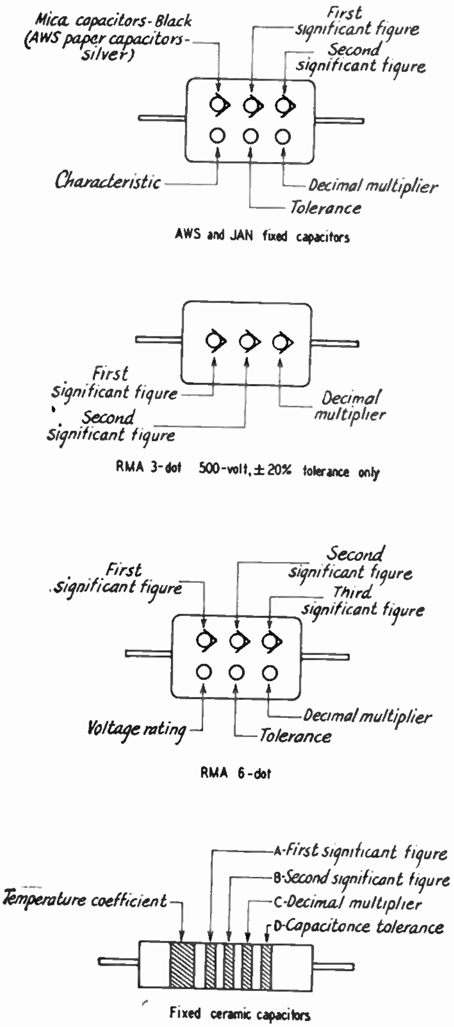


Fig. 18-5 — Color coding of fixed mica, molded paper, and tubular ceramic condensers. The color code for mica and molded paper condensers is given in Table 18-III. Table 18-IV gives the color code for tubular ceramic condensers.

± 20% tolerance only; other ratings and tolerances are covered by the 6-dot RMA code.

Examples: A condenser with a 6-dot code has the following markings: Top row, left to right, black, yellow, violet; bottom row, right to left, brown, silver, red. Since the first color in the top row is black (significant figure zero) this is the AWS code and the condenser has mica dielectric. The significant figures are 4 and 7, the decimal multiplier 10 (brown, at right of second row), so the capacitance is 470 μμfd. The tolerance is ± 10%. The final color, the characteristic, deals with temperature coefficients and methods of testing, and may be ignored.

A condenser with a 3-dot code has the following colors, left to right: brown, black, red. The significant figures are 1, 0 (10) and the multiplier is 100. The capacitance is therefore 1000 μμfd.

A condenser with a 6-dot code has the following markings: Top row, left to right, brown, black, black; bottom row, right to left, black, gold, blue. Since the first color in the top row is neither black nor silver, this is the RMA code. The significant figures are 1, 0, 0 (100) and the decimal multiplier is 1 (black). The capacitance is therefore 100 μμfd. The gold dot shows that the tolerance is ± 5% and the blue dot indicates 600-volt rating.

Ceramic Condensers

Conventional markings for ceramic condensers are shown in the lower drawing of Fig. 18-5. The colors have the meanings indicated in Table 18-IV. In practice, dots may be used instead of the *narrow* bands indicated in Fig. 18-5.

Example: A ceramic condenser has the following markings: Broad band, violet: narrow bands or dots, green, brown, black, green. The significant figures are 5, 1 (51) and the decimal multiplier is 1, so the capacitance is 51 μμfd. The temperature coefficient is — 750 parts per million per degree C., as given by the broad band, and the capacitance tolerance is ± 5%.

Fixed Composition Resistors

Composition resistors (including small wire-wound units molded in cases identical with the composition type) are color-coded as shown in Fig. 18-6. Colored bands are used on resistors having axial leads; on radial-lead resistors the colors are placed as shown in the drawing. When bands are used for color coding the body color has no significance.

Examples: A resistor of the type shown in the lower drawing of Fig. 18-6 has the following color bands: A, red; B, red; C, orange; D, no color. The significant figures are 2, 2 (22) and the decimal multiplier is 1000. The value of resistance is therefore 22,000 ohms and the tolerance is ± 20%.

A resistor of the type shown in the upper drawing has the following colors: body (A), blue; end (B), gray; dot, red; end (D), gold. The significant figures are 6, 8 (68) and the decimal multiplier is 100, so the resistance is 6800 ohms. The tolerance is ± 5%.

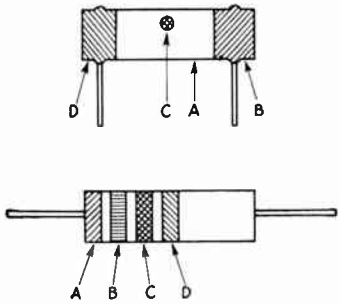
I.F. Transformers

Blue — plate lead.
Red — "B" + lead.
Green — grid (or diode) lead.
Black — grid (or diode) return.

TABLE 18-III
Resistor-Condenser Color Code

Color	*Significant Figure*	*Decimal Multiplier*	*Tolerance (%)*	*Voltage Rating**
Black	0	1	—	—
Brown	1	10	1*	100
Red	2	100	2*	200
Orange	3	1000	3*	300
Yellow	4	10,000	4*	400
Green	5	100,000	5*	500
Blue	6	1,000,000	6*	600
Violet	7	10,000,000	7*	700
Gray	8	100,000,000	8*	800
White	9	1,000,000,000	9*	900
Gold	–	0.1	5	1000
Silver	–	0.01	10	2000
No color	–	—	20	500

* Applies to condensers only.



Fixed composition resistors

Fig. 18-6 — Color coding of fixed composition resistors. The color code is given in Table 18-III. The colored areas have the following significance:
A — First significant figure of resistance in ohms.
B — Second significant figure.
C — Decimal multiplier.
D — Resistance tolerance in per cent. If no color is shown, the tolerance is ±20%.

NOTE: If the secondary of the i.f.t. is center-tapped, the second diode plate lead is green-and-black striped, and black is used for the center-tap lead.

A.F. Transformers

Blue — plate (finish) lead of primary.
Red — "B" + lead (this applies whether the primary is plain or center-tapped).
Brown — plate (start) lead on center-tapped primaries. (Blue may be used for this lead if polarity is not important.)
Green — grid (finish) lead to secondary.
Black — grid return (this applies whether the secondary is plain or center-tapped).
Yellow — grid (start) lead on center-tapped secondaries. (Green may be used for this lead if polarity is not important.)

NOTE: These markings apply also to line-to-grid and tube-to-line transformers.

Loudspeaker Voice Coils

Green — finish.
Black — start.

Loudspeaker Field Coils

Black and Red — start.
Yellow and Red — finish.
Slate and Red — tap (if any).

Power Transformers

1) Primary Leads . . . . *Black*
   If tapped:
   Common . . . . *Black*
   Tap . . . . *Black and Yellow Striped*
   Finish . . . . *Black and Red Striped*
2) High-Voltage Plate Winding . . . . *Red*
   Center-Tap . . . *Red and Yellow Striped*
3) Rectifier Filament Winding . . . . *Yellow*
   Center-Tap . . *Yellow and Blue Striped*
4) Filament Winding No. 1 . . . . *Green*
   Center-Tap . . *Green and Yellow Striped*
5) Filament Winding No. 2 . . . . *Brown*
   Center-Tap . *Brown and Yellow Striped*
6) Filament Winding No. 3 . . . . *Slate*
   Center-Tap . . . *Slate and Yellow Striped*

TABLE 18-IV
Color Code for Ceramic Condensers

Color	*Significant Figure*	*Decimal Multiplier*	*Capacitance Tolerance*		*Temp. Coeff. p.p.m./deg. C.*
			More than 10 μμfd. (in %)	*Less than 10 μμfd. (in μμfd.)*	
Black	0	1	± 20	2.0	0
Brown	1	10	± 1		—30
Red	2	100	± 2		—80
Orange	3	1000			—150
Yellow	4				—220
Green	5		± 5	0.5	—330
Blue	6				—470
Violet	7				—750
Gray	8	0.01		0.25	30
White	9	0.1	± 10	1.0	500

Eliminating Broadcast Interference

It is your duty as an amateur to make sure that the operation of your station does not interfere with broadcasting or other radio services because of any shortcomings in your equipment. Failure to observe this rule may lead to curtailed operating privileges — a situation that is easily avoidable if you build and adjust your transmitter according to good practice.

However, there is a larger obligation — to eliminate broadcast interference to the greatest possible extent even when your own transmitter is not at fault. The institution of amateur radio cannot continue to flourish in the face of ill feeling on the part of a large segment of the general public — ill feeling that is only too readily generated if the public's favorite radio programs are broken up by amateur transmissions. It is no exaggeration to say that the future of amateur radio depends in large part on the efforts you exert now to make it possible for your neighbors to continue to enjoy their radio reception while you pursue your transmitting activities. It is unfortunately true that most interference to broadcasting is directly the fault of present-day broadcast-receiver construction. Nevertheless, the amateur can and should help to alleviate interference even though the responsibility for it does not lie with him.

The regulation of the Federal Communications Commission covering interference to broadcasting is quoted below:

§ 12.152. *Restricted operation.* (a) If the operation of an amateur station causes general interference to the reception of transmissions from stations operating in the domestic broadcast service when receivers of good engineering design including adequate selectivity characteristics are used to receive such transmissions and this fact is made known to the amateur station licensee, the amateur station shall not be operated during the hours from 8 o'clock P.M. to 10:30 P.M., local time, and on Sunday for the additional period from 10:30 A.M. until 1 P.M., local time, upon the frequency or frequencies used when the interference is created. (b) In general, such steps as may be necessary to minimize interference to stations operating in other services may be required after investigation by the Commission.

FCC recognizes the fact that much of the interference that occurs is because receivers are not capable of rejecting signals far outside the frequency band to which the receiver is tuned. That is why the phrases "general interference" and "receivers of good engineering design including adequate selectivity characteristics" are used in Section 12.152. "Quiet hours" are not imposed unless it is shown that the interference is actually the fault of the transmitter.

Once you have determined that your transmitter is free from parasitic oscillations, spurious radiations, key clicks and modulation splatter, you can tackle the BCI problem with a clear conscience and the firm conviction that the answer is to be found in the b.c. receiver. Be sure your transmitter is clean *first*. From then on you have a twofold job: convincing the owner of the receiver that his set is at fault (not always the easiest thing in the world, especially if the receiver is fairly new), and finding out just why the interference occurs. The first is almost wholly a matter of using the right approach: you may need all the tact at your command to convince him that you know what you're talking about and are sincerely trying to help. His natural tendency, as one with no technical knowledge of radio at all, will be to blame you because you're coming in on the broadcast band where you obviously don't belong. You may have to overcome the suspicion that everything you say about his receiver is just so much camouflage to cover up something wrong with your transmitter.

In brief, to be successful in eliminating BCI you have got to win the listener's coöperation.

● GETTING LISTENER COOPERATION

The battle is 75 per cent won when you've earned the listener's confidence in your technical ability and your sincerity in wanting to clear up interference. Here are a few pointers on how to go about it.

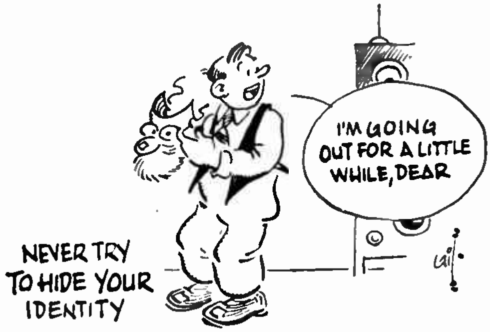
Clean House First

We've said above that the first obligation of every amateur is to clean up his transmitter so it has no radiations outside the bands assigned for amateur use. Even then, you'll probably find that you have a BCI problem in your own house.

So clean up your household BCI first! It is always convincing if you can say — and demonstrate — that you do not interfere with broadcast reception in your own home.

Don't Hide Your Identity

If a listener thinks that you are "trying to get away with something" he will not only be unwilling to coöperate, but may be actively hostile. As a general rule, whenever you change location, or mode of transmission, or increase power, or put up a new antenna, check with your neighbors to make sure that they are not



experiencing interference. Announce your presence and conduct occasional tests on the air, requesting anyone whose reception is being spoiled to let you know about it so that you may take steps to eliminate the trouble.

Act Promptly

Do something to show the listener that you are concerned for his welfare as soon as a complaint is received. The average person will tolerate a limited amount of interference, but no one can be expected to put up with frequent and extended interruption of his listening pleasure. The sooner you take steps to eliminate the interference, the more agreeable the listener will be; the longer he has to wait for you, the less willing he will be to coöperate.

Present Your Story Tactfully

Put yourself in the listener's place. He has a right, he believes, to interference-free reception of the broadcast programs he likes. When you interfere, his natural reaction is to assume that you are the one at fault. When you call on him, explain that you do not operate on the frequencies to which he wants to listen, and the real trouble is that you and he happen to be located so close to each other. Explain to him that there are thousands of stations operating simultaneously, all the time, and that the problem of rejecting all but the one he happens to want to hear is one of receiver design. Point out that the average broadcast receiver is made to sell as cheaply as possible, and that features that would prevent interference from near-by stations are left out.

It should be explained to the listener that if it is simply the presence of your strong signal on his receiving antenna that causes the difficulty, the situation can be cleared up by a wavetrap. In other cases the wiring of the receiver itself is picking up your signal, and such cases can be cured only by suppressing this unwanted pick-up in the receiver itself; in other words, some modifications will have to be made in the receiver if he is to expect interference-free reception.

Arrange for Tests

Most listeners are not very competent observers of the various aspects of interference. If at all possible, enlist the help of another amateur and have him operate your transmitter while you see what happens at the affected broadcast set. You can then determine for yourself where the trouble is most likely to be.

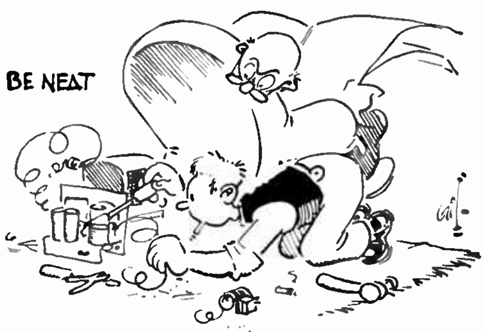
It is a good idea to take along a wavetrap when you arrange such a test. If the receiver is one having an external antenna, it may be possible to cure the interference then and there.

Avoid Working on the Receiver

If your tests show that the fault has to be remedied in the receiver itself, *do not offer to work on the receiver*. It is not your fault that the receiver design is defective. Recommend that the work be done by a reliable serviceman, and offer to advise the latter as to the cause and cure if necessary.

It is inadvisable to tackle broadcast receivers, particularly the midget varieties, unless you have had experience working on them. In any event, if you do work on the receiver yourself the chances are that if anything goes wrong later on you'll be blamed for it. Explain that, while you may be technically competent to make the necessary modifications, radio servicing is best left to those who specialize in it, and that you are sure he, the owner, will prefer to have the work done by someone whom he can hold responsible.

If the owner of the receiver obviously prefers to have you make the modifications, do so only with the understanding that it is purely as a favor and because you are anxious to coöperate. Make him understand, with as much tact as possible, that the responsibility for the interference does not lie with you (your transmitter having previously been checked and found OK); if the receiver responds to fre-



quencies to which it is not tuned that is a defect in its design. You also have no obligation to pay for having the receiver modified. If you do the work yourself you should not make any charge, of course. In that event, insist that you must take the receiver to your own shop in order to work on it properly; you will be able to tell immediately whether the changes you make effect an improvement and therefore can work more rapidly and conveniently — and without turning the owner's living room into a repair shop. If it is necessary to do some work in the listener's home, *be neat* in the work you do. Remember, the listener's living room cannot be treated in the same manner you would treat your own ham shack!

In General

In this "public relations" phase of the problem a great deal depends on your own attitude. Most people will be willing to meet you half way, particularly when the interference is not of long standing, if you as a person make a good impression. Your personal appearance is important. So is what you say about the receiver. A display of lofty technical superiority is more likely to generate resentment than coöperation. Above all, don't make remarks on the air about "bum broadcast receivers" and "cheap midgets." No one takes kindly to hearing his possessions publicly derided. If you discuss your BCI problems on the air, do it in a constructive way — one calculated to increase listener coöperation, not destroy it.

RADIO-CLUB BCI COMMITTEES

Organized amateur radio clubs can do a lot to pave the way toward coöperation between individual amateurs and the broadcast listeners. Most clubs maintain interference committees charged with handling both the public relations and the technical aspects of BCI. Through such committees, technical assistance is made available to all members of the club so that those less qualified can have the benefit of the experience of others. The committee should also maintain contact with the local radio servicemen, supplying them with information and technical assistance whenever possible. The committee can maintain valuable contacts with the local newspapers, broadcast stations and other authorities to provide the right kind of publicity for the efforts of individuals or groups who are trying to clear up BCI problems.

League Aids

The Communications Department of ARRL, as one of its services to affiliated clubs, has prepared material suggesting various ways in which local clubs can form interference committees, and methods by which such groups can function efficiently for the good of all concerned. This material is available to affiliated clubs on request, addressed to ARRL headquarters.

Causes and Cure of BCI

There are no magic cures for all cases of interference to standard AM broadcasting. The great number of different types of broadcast receivers makes it necessary to tailor the remedy to the specific set. However, interference does usually fall into one or more rather well-defined categories. A knowledge of the general types of interference and the methods required to eliminate it will lead to a rapid appraisal of the situation and will avoid much cut-and-try in finding a cure.

Transmitter Defects

Out-of-band radiation is something that must be cured at the transmitter. Parasitic oscillations are a frequently unsuspected source of such radiations, and no transmitter can be considered satisfactory until it has been thoroughly checked for both low- and high-frequency parasitics. Very often parasitics show up only as transients, causing key clicks in c.w. transmitters and "splashes" or "burps" on modulation peaks in AM transmitters. Methods for detecting and eliminating parasitics are discussed in Chapter Six.

In c.w. transmitters the sharp make and break that occurs with unfiltered keying causes transients that, in theory, contain frequency components through the entire radio spectrum. Practically, these transients do not have very much amplitude at frequencies very far away from the transmitting frequency. Nevertheless they are often strong enough in the immediate vicinity of the transmitter to cause serious interference to broadcast reception. Key clicks can be eliminated by the methods detailed in Chapter Eight.

A distinction must be made between clicks generated in the transmitter itself and those set up by the mere opening and closing of the key contacts when current is flowing. The latter are of the same nature as the clicks heard in a receiver when a wall switch is thrown to turn a light on or off, and may be more troublesome nearby than the clicks that actually go out on the signal. A filter for eliminating them usually has to be installed as close as possible to the key contacts.

Overmodulation in AM 'phone transmitters generates transients similar to key clicks. It can be prevented either by using automatic systems for limiting the modulation to 100 per cent, or by continuously monitoring the modulation. Methods for both are described in Chapter Nine. In this connection, the term "overmodulation" means any type of non-linear modulation that results from overloading or inadequate design. This can occur even though the actual modulation percentage is less than 100.

BCI is frequently made worse by radiation from the transmitter, power wiring, or the r.f. transmission line. This is because the signal causing the interference, in such cases, is radiated from wiring that is nearer the broadcast receiver than the antenna itself. In such cases much depends on the method used to couple the transmitter to the antenna, a subject that is discussed in Chapter Ten. If it is at all possible, too, the antenna itself should be placed so that it is not in close proximity to house wiring, telephone and power lines, and similar conductors.

Image and Oscillator-Harmonic Responses

Relatively few superhet broadcast receivers have any r.f. amplification preceding the mixer, so that the selectivity at the signal frequency is not especially high (the i.f. amplifier provides most of the working selectivity). The result is that strong signals from near-by transmitters, even though the transmitting frequency is far removed from the broadcast band, can force themselves to the mixer grid. They will normally be eliminated by the i.f. selectivity, except in cases where the transmitter frequency is the image of the broadcast signal to which the receiver is tuned, or when the transmitter frequency is so related to a harmonic of the broadcast receiver's local oscillator as to produce a beat at the intermediate frequency.

These image and oscillator-harmonic responses tune in and out on the broadcast receiver dial just like a broadcast signal, except that in the case of harmonic response the tuning rate is more rapid. Since most receivers use an intermediate frequency in the neighborhood of 450 kc., the interference is a true image only when the amateur transmitting frequency is in the 1750-kc. band. Oscillator-harmonic responses occur from 3.5- and 7-Mc. transmissions, and sometimes even from higher frequencies.

Regardless of whether the interference is caused by either an image or by harmonic response, the problem is to reduce the amplitude of the amateur signal in the front end of the b.c. receiver. If the receiver uses an external antenna a wavetrap at the receiver antenna terminals may help. It may also be helpful to reduce the length of the receiving antenna — and particularly to avoid a length that might be near resonance at the transmitter frequency — or to change its direction with respect to the transmitting antenna. If the signal is being picked up by the antenna it will disappear when the antenna is disconnected. If it is still present under these circumstances the pick-up is in the set wiring or the power circuits. A line filter may be tried for the latter. Pick-up on the set wiring can only be cured by installing some shielding around the r.f. circuits. Copper window screening cut and fitted to size will usually do the trick.

Since images and harmonic responses occur at definite frequencies on the receiver dial, it is always possible to choose an operating frequency that will not give such a response on top of the broadcast stations that are favored in the vicinity. While your signal may still be heard when the receiver is tuned off the local stations, it will at least not interfere with program reception.

Cross-Talk

With some of the older receivers, particularly of the nonsuperheterodyne type, interference occurs only when the receiver is tuned to a strong broadcast signal and disappears between stations. This is cross-modulation, a result of rectification in one of the early stages of the receiver. It is not so likely to occur in more modern sets using a remote-cut-off tube in the antenna stage.

One remedy is to install remote-cut-off tubes in the r.f. stages and put in an a.v.c. circuit. However, this is a major operation and frequently is not practicable. The remaining thing is to reduce the strength of the amateur signal at the grid of the first tube in the receiver. Wavetraps, a smaller antenna, and a different antenna position should be tried. Additional shielding about the r.f. circuits also will sometimes effect an improvement.

Blanketing

"Blanketing" is a form of interference that partially or completely masks reception, no matter where the broadcast receiver is tuned. Each time the carrier is thrown on, whether by keying or for modulation, the program disappears or is greatly reduced in amplitude. Amplitude modulation in such a case is usually distorted rather severely.

When the transmitter is operated on the lower frequencies this type of interference occurs only when the receiver and transmitter are very close together. It is the result of simple overloading of the receiver by the very strong field in the vicinity of the transmitting antenna. It occurs principally on receivers using external antennas (as contrasted with a built-in loop), and can be reduced by the steps recommended above; i.e., using a short receiving antenna, repositioning the antenna with respect to the transmitting antenna so the pick-up is reduced, or using wavetraps and line filters.

When the transmitter is operated on 28 Mc. or v.h.f. "blanketing" occurs rather rarely, and then only when the transmitting and receiving installations are located exceptionally close together.

Audio-Circuit Rectification

The most frequent cause of interference from operation at the higher frequencies is from rectification of a signal that by one means or another gets into the audio system of the receiver. In the milder cases an amplitude-modulated signal will be heard with reasonably good quality, but is not tunable — that is, it is present no matter what the frequency to which the receiver dial is set. An unmodulated carrier may have no observable effect in such cases beyond causing a little hum. However, if the signal is very strong there will be a reduction of the audio output level of the receiver whenever the carrier is thrown on. This causes an annoying "jumping" of the program when the interfering signal is keyed. With 'phone transmission the change in audio level is not so objectionable because it occurs at less frequent intervals. Also, ordinary rectification gives no audio output from a frequency-modulated signal, so the interference can be made almost completely unnoticeable if FM or PM is used instead of AM.

Interference of this type is most prevalent in a.c.-d.c. receivers. The pick-up may occur in the audio-circuit wiring or the interfering signal may get into the audio circuits by way of the line cord. Power-line pick-up can be treated by means of line filters, but pick-up in the receiver wiring requires individual attention. Remedies that have been found successful are described in the sections following.

CHECKING AND CURING BCI

When a case of broadcast interference comes to your attention, set a definite time to conduct tests and then prepare to do the job as expeditiously as possible. Provide yourself with one or two wavetraps and line filters, since they can be tried immediately without getting into the receiver. As suggested before, get another amateur to operate your transmitter while you do the actual observing and testing at the listener's receiver. The procedure outlined below will save time in getting at the source of the trouble and in satisfactorily eliminating it.

1) Determine whether the interference is tunable or not. This will usually indicate the methods required for elimination of the trouble, as it will show which of the general types of interference discussed above is present. In severe cases it is possible that two or more types will be present at the same time, and steps will be necessary to eliminate each type.

2) If the set has an external antenna, disconnect it and turn the volume control up full. If the interference is no longer present, it is merely necessary to prevent the r.f. appearing on the antenna from entering the set. If wavetraps reduce the amplitude of the interfering signal but do not eliminate it entirely, try a short piece of wire as a receiving antenna. Alternatively, the antenna may be relocated. It should be placed as far as possible from the transmitting antenna, and should run at right angles to it to minimize coupling.

If the interference persists after the antenna is disconnected, the search is narrowed to an investigation of whether the signal is coming in on the power lines, or is being picked up directly on the receiver wiring.

3) Check for power-line interference by using a sensitive wavemeter such as that described in Chapter Sixteen of this *Handbook* to probe along the a.c. cord that connects the set to the power source. Checks should be made at the transmitter frequency, and also at harmonic frequencies. If r.f. is detected in the line, by-pass both sides of the a.c. line to ground with 0.005-μfd. mica condensers at the point where the line cord enters the set. (A simple plug-and-socket adapter can be made up for this purpose before visiting the listener.) If this does not completely eliminate the interference, try a line filter designed for the operating frequency.

4) If it is evident that the interference is being picked up on the receiver wiring, explain the situation to the owner and tell him that the exact cause cannot be determined without removing the chassis from the cabinet, and that, in any event, the receiver will have to be modified somewhat if the interference is to be eliminated. As suggested before, recommend that the actual work be done by a radio serviceman. Offer to check into the cause yourself, if he wishes and will allow you to take the set to your shop (with the understanding that you will not make any changes in the receiver without his express permission) so the serviceman can be told what needs to be done.

5) In the event that the owner allows you to take the receiver, set it up near your transmitter and check to see if the amplitude of the interfering signal is changed by various settings of the receiver volume control. If the volume of the interference changes with changes in the volume control, the r.f. is entering the set *ahead* of the volume control. If it is unaffected by the volume control, it is getting into the audio stages at a point following the volume control.

6) Pin the source down, if it is ahead of the volume control, by removing one tube at a time until one is found that kills the interference when it is removed. In sets using series-connected filaments, this will be possible only if a tube of equal heater rating, and with all but the heater pins clipped off, is *substituted* for the tube.

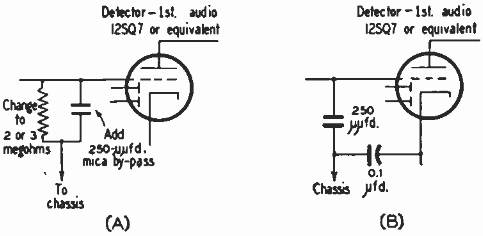


Fig. 19-1 — Two methods of eliminating r.f. from the grid of a combined detector/first-audio stage. At A, the value of the grid leak is reduced to 2 or 3 megohms, and a mica by-pass condenser is added. At B, both grid and cathode are by-passed.

7) Determine which element (or elements) of the tube is picking up the interference by touching each tube pin with a test lead about three feet long. The lead, acting as an antenna, will cause the interference to increase when it is placed on a tube pin that is contributing to the interference. Once the sensitive points have been determined, the trouble can be eliminated by shielding the leads connected to the tube element that is affected, and by shielding

the tube itself. Grid leads are the principal offenders, especially the long leads that run from a tube cap to a tuning condenser, and it may be necessary to shield several parts of the set before the interference is eliminated.

8) If the pick-up is found to be in the audio system — as is the case in many sets, especially when the transmitter is operating at 28 Mc. or higher — it can be eliminated by one or another of the methods shown in Figs. 19-1 and 19-2. Fig. 19-1A is a method that has proved successful with many a.c.-d.c. receivers. The value of the grid leak in the combined detector/first-audio tube (usually a 12SQ7 or its equivalent) is reduced to 2 or 3 megohms. The grid is then by-passed for r.f. with a 250-$\mu\mu$fd. mica condenser. Fig. 19-1B is a similar method. A third method that has worked in

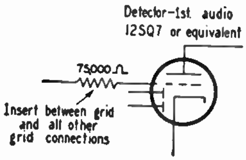


Fig. 19-2 — Using a 75,000-ohm resistor to form a low-pass filter with the tube capacitance. The resistor must be mounted at the tube pin, between the grid and all other grid connections.

a.c.-d.c. receivers requires only that the heater of the detector/first-audio stage be by-passed to ground with a 0.001-μfd. condenser. The method shown in Fig. 19-2 uses a 75,000-ohm ½-watt resistor to form, with the tube capacitance, a low-pass filter. The resistor is connected between the grid pin of the audio stage and *all other* wires connected to the grid. In all cases, both sides of the a.c. line should be by-passed to chassis with 0.001- to 0.01-μfd. condensers.

Wavetraps and A.C. Line Filters

A wavetrap consists of a parallel-tuned circuit that is connected in series with the broadcast antenna and the antenna post of the receiver. It should be designed to resonate at the frequency of the interfering signal. The circuit of a simple trap is shown in Fig. 19-3. If interference results from operation in more than one amateur band several traps may be connected

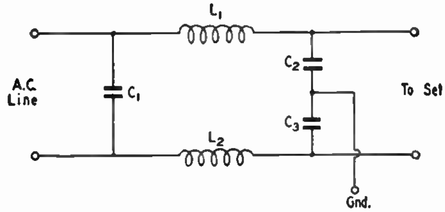


Fig. 19-4 — A.c. line filter for receivers. The values of C_1, C_2 and C_3 are not generally critical; capacitances from 0.001 to 0.01 μfd. can be used. L_1 and L_2 can be a 2-inch winding of No. 18 enameled wire on a half-inch diameter form.

in series, each tuned to the center of one of the bands in which operation is contemplated. To adjust the wavetrap, have another licensed amateur operate the transmitter while you tune the trap for maximum attenuation of the interference. The trap should be connected to the broadcast receiver and the normal receiving antenna should be connected in series with the trap, as shown in the figure.

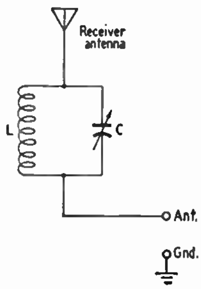


Fig. 19-3 — A simple wavetrap circuit. L and C must resonate at the frequency of the interfering signal. Suitable constants are tabulated below.

Band	*C*	*L*			
3.5	140 $\mu\mu$fd.	16 μh.,	32 turns	#22, 1" diam.,	1" long
7	100 $\mu\mu$fd.	6	19	#22, 1"	1"
14	50 $\mu\mu$fd.	3.5	14	#18, 1"	1"
21	35 $\mu\mu$fd.	2.2	12	#18, 1"	1"
28	25 $\mu\mu$fd.	1.5	9	#18, 1"	1"

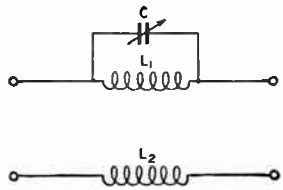


Fig. 19-5 — Resonant filter for the a.c. line. A single condenser tunes both L_1 and L_2, which are unity-coupled, one wound on top of the other. Constants for amateur bands are tabulated below.

Band	*C*	*L_1 - L_2*
3.5	140 + 150 (fixed)	25 t. No. 18, 1¼" dia. × 2⅜" long
7	140 $\mu\mu$fd.	18 t. No. 18, 1¼" dia. × 2⅜" long
14	100 $\mu\mu$fd.	12 t. No. 18, 1¼" dia. × 2⅜" long
21	50 $\mu\mu$fd.	10 t. No. 18, 1¼" dia. × 2⅜" long
10	25 $\mu\mu$fd.	9 t. No. 18, 1½" dia. × 2⅜" long

D.c.c. wire is recommended for all coils.

A common form of a.c. line filter is shown in Fig. 19-4. This type of filter will usually do some good if the signal is being picked up on the house wiring and transferred to the set by way of the line cord. The values used for the coils and condensers are in general not critical. The effectiveness of the filter will depend considerably on the ground connection used, and it may be necessary to try grounding to several different possible ground connections to secure the best results. A filter of this type will usually not be very helpful if the signal is being picked up on the line cord itself, which may be the case when the transmitter is on v.h.f. In such a case it should be installed inside the receiver chassis and grounded to the chassis at the point where the line cord enters.

The tuned filter shown in Fig. 19-5 is often more effective than the untuned type when only one frequency needs to be eliminated. After installation, the condenser is simply adjusted to reduce the interference to the greatest possible extent.

It is advisable to mount either type of filter in a small shielding box, both to prevent pick-up in the filter itself and to make it less conspicuous when it has to be installed in a listener's home.

Interference with Television

Interference with reception of television signals presents a more difficult problem than interference with ordinary AM broadcasting. In the latter case it is comparatively easy to clean up a transmitter so that it will have no spurious radiations in the broadcast band. Clearing up interference difficulties then becomes a matter of overcoming deficiencies in the selectivity of the broadcast receiver.

In the case of television reception similar receiver deficiencies exist, and must be treated by methods similar to those used for low-frequency broadcasting. However, a more serious situation for the amateur arises because harmonics of his transmitting frequency fall in many of the television channels. The relationship between television channels and harmonics of amateur bands from 14 through 28 Mc. is shown in Fig. 19-6. Harmonics of the 7- and 3.5-Mc. bands are not shown because they fall in every television channel. Also, the harmonics above 54 Mc. from these bands are of such high order that they are usually rather low in amplitude. They are not, however, too weak to interfere if the television receiver is quite close to the amateur transmitter.

Low-order harmonics — up to about the fourth or fifth — are usually the most difficult to eliminate. The degree of harmonic suppression required is very great, particularly when the television receiver is nearby and the signals from television stations are weak. Effective harmonic suppression has three separate phases:

1) Reducing the amplitude of harmonics generated in the transmitter. This is a matter of circuit design and operating conditions.

2) Preventing stray radiation from the transmitter and from associated wiring. This requires adequate shielding and filtering of all circuits and leads from which radiation can take place.

3) Preventing harmonics from being fed into the antenna.

Methods for reducing the amplitude of generated harmonics and for preventing stray radiation are detailed in Chapter Six. Chapter Ten gives information on preventing harmonics from reaching the transmitting antenna.

Checking Procedure

Interference with television may be caused either by the fundamental output of the transmitter (overloading) or by harmonics that fall in the TV channel. Except possibly in the case of transmitters working at 3.5 and 1.75 Mc., it is safe to assume that harmonics are at least partly responsible. Receiver overloading because of the strong fundamental radiation from the transmitter occurs only when the TV receiver is very close — within a couple of hundred feet of the transmitter.

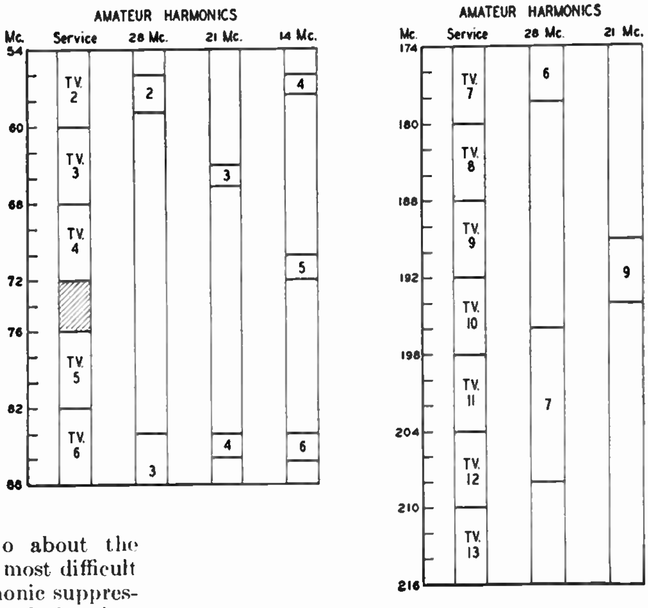


Fig. 19-6 — Relationship of amateur-band harmonics to TV channels. Harmonic interference is most likely to be serious in the low-channel group (54 to 88 Mc.).

Interference from the fundamental frequency of the transmitter usually can be identified by the fact that it occurs on all channels, whether or not they have a direct harmonic relationship with the fundamental. This type of interference is considered in detail in a subsequent section.

True harmonic interference occurs only on TV channels in which the transmitter harmonics fall. To reduce it effectively it is necessary first to determine the cause of the harmonic radiation. Measures taken to reduce the amplitude of harmonics fed to the transmitting antenna will have little effect if the principal radiation is coming from the transmitter itself, and vice versa. Before any corrective measures are tried, therefore, the antenna should be disconnected from the transmitter and replaced by a suitable dummy antenna. If the interference is still present, the transmitter itself must be made radiation-free by the methods outlined in Chapter Six. It is not worth while to do any further testing with the regular transmitting antenna until the set radiation is entirely eliminated. But once the transmitter can be operated into a dummy antenna without causing interference, it is then certain that harmonic interference caused when the regular antenna is connected will respond to treatment by the methods given in Chapter Ten.

Testing with a dummy antenna also practically eliminates fundamental interference, except possibly when the transmitter and TV receiver are within a few feet of each other. If, on reconnecting the antenna after the set radiation is eliminated, there is still interference on channels not in harmonic relation to the transmitting frequency, the measures described in the next section should be applied to the TV receiver before proceeding further with harmonic reduction at the transmitter.

RECEIVER DEFICIENCIES

Spurious responses because of receiver inadequacies are particularly likely to occur when the receiver and transmitter are quite close. They usually result from the fact that the strong fundamental-frequency signal from the transmitter overloads some circuit in the receiver.

Many television receivers have "front ends" that are inherently unselective and not well balanced — that is, they will give strong response to parallel currents on the receiving transmission line. Usually, the transmission line picks up a great deal more energy from a near-by transmitter than the television receiving antenna itself, causing parallel currents that should be, but are not, rejected by the receiver's input circuit. A strong signal that overloads the first or second stages in the receiver will cause the receiver itself to generate harmonics that fall in the television channels. This situation can be improved by using shielded transmission line — coax or, in the balanced form, "twinax" — on the receiving installation. For best results the line should terminate in a coax fitting on the receiver chassis, but if this is not possible the shield should be grounded to the chassis right at the antenna terminals.

The use of shielded transmission line also will be helpful in reducing response to harmonics actually being radiated from the transmitter or transmitting antenna. In most receiving installations the transmission line is very much longer than the antenna itself, and is consequently far more exposed to the harmonic fields from the transmitter. Much of the harmonic pick-up, therefore, is on the receiving transmission line when the transmitter and receiver are quite close together. Shielded line, plus relocation of either the transmitting or receiving antenna to take advantage of directive effects, often will result in reducing the harmonic pick-up to a level that does not interfere with reception.

Many television receivers do not have enough isolation between the antenna and intermediate-frequency circuits. As a result, signals that fall in or near the intermediate-frequency passband (roughly 22 to 27 Mc. in most current receivers) will cause interference either to the picture or to the sound. If the receiver and transmitter are very close a complete cure may not be possible without shielding the receiver's i.f. circuits. I.f. interference is particularly likely from the 21-Mc. band when the receiver has its sound i.f. channel centered at or near 21.25 Mc. Realigning the receiver to a somewhat higher frequency (sound channel at 21.9 Mc.) usually will cure this type of interference.

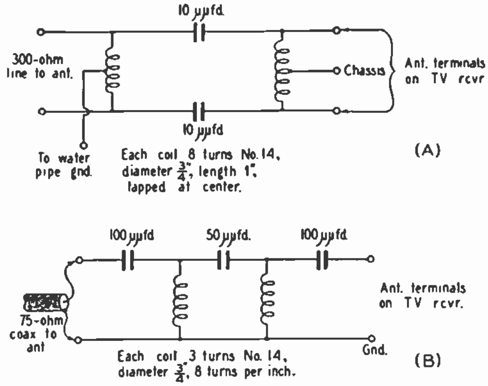


Fig. 19-7 — High-pass filters for installation at the TV receiver antenna terminals. A — balanced filter for 300-ohm line, B — for 75-ohm coaxial line. *Important:* Do not use a direct ground on an a.c.-d.c. chassis. Ground through a 0.001-μfd. mica condenser.

If the fundamental signal is getting into the receiver by way of the line cord a line filter such as that shown in Fig. 19-4 will help. To be most effective it should be installed inside the receiver chassis at the point where the cord enters, making the ground connections directly to chassis at this point. It may not be so helpful if placed between the line plug and the wall socket unless the r.f. is actually picked up on the house wiring rather than on the line cord itself.

In cases where the fundamental r.f. is known to be reaching the receiver through the antenna and transmission line, it can be prevented from doing much harm by installing a high-pass filter at the receiver's antenna terminals. Circuits that have proved effective are shown in Figs. 19-7 and 19-8. Fig. 19-8 has one more

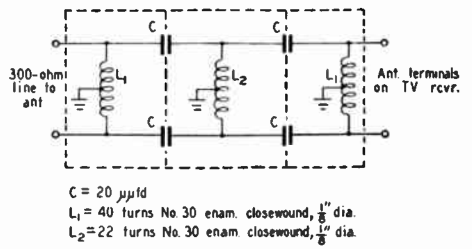


Fig. 19-8 — Another type of high-pass filter for 300-ohm line. The coils may be wound on 1/8-inch diameter plastic knitting needles. *Important:* Do not use a direct ground on an a.c.-d.c. chassis. Ground through a 0.001-μfd. mica condenser.

section than the filters of Fig. 19-7 and as a consequence has somewhat better cut-off characteristics. All the circuits given are designed to have little or no effect on the TV signals but will attenuate all signals lower in frequency than about 40 Mc. These filters preferably should be constructed in some sort of shielding container, although shielding is not always necessary. The dashed lines in Fig. 19-8 show how individual filter coils can be shielded from each other. The condensers can be ceramic units centered in holes in the partitions that separate the coils.

High-pass filters designed for this purpose are available commercially at moderate prices. In this connection, it should be understood by all parties concerned that while an amateur is responsible for *harmonic* radiation from his transmitter, it is no part of his responsibility to pay for or install filters, wavetraps, etc., that may be required at the receiver to prevent interference caused by his *fundamental* frequency. It is a good idea for the amateur to have a high-pass filter that can be tried on the receiver when interference exists. If trial shows it to be effective, the reason why it works should be carefully explained to the set owner, who should then be advised to get in touch with the organization from which he purchased the receiver or which services it, to make arrangements for proper installation. The question of cost is one to be settled between the set owner and the organization with which he deals.

Wavetraps may be used instead of high-pass filters. If the receiver has a balanced (300-ohm) transmission line a trap should be used in each line wire. They may be constructed from the data in Fig. 19-3. When properly tuned, wavetraps will greatly attenuate the fundamental signal but suffer the disadvantage, as compared with a high-pass filter, that they must be retuned if the transmitter frequency is moved. They are of course of no value in rejecting a frequency to which they cannot be tuned, and therefore usually are good only for one amateur band.

Another type of interference, wholly attributable to lack of receiver selectivity, occurs from operation in the 50-Mc. band. A strong 50-Mc. signal on the receiving antenna will overload the receiver, particularly when the receiver is tuned to Channel 2. Wavetraps tuned to the frequency of the interfering signal, installed at the antenna input terminals of the receiver, will help reduce this type of interference. It is also helpful to work at the low-frequency end of the 50-Mc. band, since this frequency is farthest removed from Channel 2. Shielding of the receiver's r.f. circuits also may be necessary.

Operating a Station

The enjoyment of our hobby usually comes from the operation of our station once we have finished its construction. Upon the *station* and its *operation* depend the communication records that are made.

An operator with a slow, steady, clean-cut method of sending has a big advantage over the poor operator. Good sending is partly a matter of practice but patience and judgment are just as important qualities of an operator as a good "fist." The technique of speaking in connected thoughts and phrases is equally important for the operator who uses voice.

OPERATING COURTESY AND TOLERANCE

Normal operating interests in amateur radio vary considerably. Some prefer to rag-chew, others handle traffic, others work DX, others concentrate on working certain areas, countries or states, still others get on for an occasional contact only to check a new rig or antenna.

Interference is one of the things we amateurs have to live with. However, we can conduct our operating in a way designed to alleviate it as much as possible. *Before putting the transmitter on the air, listen on your own frequency.* If you hear stations engaged in communication on that frequency, stand by until you are sure no interference will be caused by your operations, or *shift to another frequency.* No amateur or any group of amateurs has any *exclusive* claim to any frequency in any band. We must work together, each respecting the rights of others. Remember, those other chaps can cause you as much interference as you cause them, sometimes more! Where a VFO is used it is not necessary to stick to a single operating frequency though it is well to have one or two preferred and alternate frequencies. It has become general operating procedure these days to work stations on or near your own frequency. This practice will automatically assist in reducing interference.

C.W. PROCEDURE

The best operators, *both* those using voice and c.w., observe certain procedures developed from experience and regarded as "standard practice."

1) *Calls.* Calling stations may call efficiently by transmitting the call signal of the station called three times, the letters DE, followed by one's own station call sent three times. (Short calls with frequent "breaks" to listen have proved to be the best method.) Repeating the call of the station called five times and signing not more than twice (repeating not more than three times) has proved excellent practice, thus: WØBY WØBY WØBY WØBY WØBY DE W1AW W1AW [etc.] $\overline{AR}$.

CQ. The general-inquiry call (CQ) should be sent not more than five times without interspersing one's station identification. The length of repeated calls is carefully limited in intelligent amateur operating. (CQ is not to be used when testing or when the sender is not expecting or looking for an answer. Never send a CQ "blind." Always listen on the frequency first.)

The directional CQ: To reduce the number of useless answers and lessen QRM, every CQ call should be made informative when possible.

> *Examples:* A United States station looking for any Hawaiian amateur calls: CQ KH6 CQ KH6 CQ KH6 DE W4IA W4IA W4IA K. A Western station with traffic for the East Coast when looking for an intermediate relay station calls: CQ EAST CQ EAST CQ EAST DE W5IGW W5IGW W5IGW K. A station with messages for points in Massachusetts calls: CQ MASS CQ MASS CQ MASS DE W7CZY W7CZY W7CZY K. In each example indicated it is understood that the combination used is repeated three times.

Hams who do not raise stations readily may find that their sending is poor, their calls ill-timed or judgment in error. When conditions are right to bring in signals from the desired locality, you can call them. Reasonably short calls, with appropriate and brief breaks to listen, will raise stations with minimum time and trouble.

2) *Answering a Call:* Call three times (or less); send DE; sign three times (or less); after contact is established decrease the use of the call signals of both stations to *once or twice.* When a station receives a call without being certain that the call is intended for it QRZ? may be used. It means "By whom am I being called?" QRZ should not be used in place of CQ.

3) *Ending Signals and Sign-Off:* The proper use of $\overline{AR}$, K, $\overline{KN}$, $\overline{SK}$ and CL ending signals is as follows:

$\overline{AR}$ — End of transmission. Recommended after call to a specific station before contact has been established.

> *Example:* W6ABC W6ABC W6ABC DE W9LMN W9LMN W9LMN $\overline{AR}$. Also at the end of transmission of a radiogram, immediately following the signature, preceding identification.

K — Go ahead (any station). Recommended after CQ and at the end of each transmission during QSO when there is no objection to others breaking in.

> *Example:* CQ CQ CQ DE W1ABC W1ABC W1ABC K or W9XYZ DE W1ABC K.

$\overline{KN}$ — Go ahead (specific station), all others keep out. Recommended at the end of each transmission during a QSO, or after a call, when calls from other stations are not desired and will not be answered.

> *Example:* W4FGH DE XU6GRL $\overline{KN}$.

$\overline{SK}$ — End of QSO. Recommended before signing *last* transmission at end of a QSO.

> *Example:* . . . . $\overline{SK}$ W8LMN DE W5BCD.

CL — I am closing station. Recommended when a station is going off the air, to indicate that it will not listen for any further calls.

> *Example:* . . . . $\overline{SK}$ W7HIJ DE W2JKL CL.

4) *Test signals* to permit another station to adjust receiving equipment may consist of a series of Vs with the call signal of the transmitting station at frequent intervals. Remember that a test signal can be a totally unwarranted cause of QRM, and *always listen first* to find a clear spot if possible.

5) *Receipting* for conversation or traffic: Never send acknowledgment until the transmission has been entirely received. "R" means "All right, OK, I understand *completely*." Use R *only* when *all* is received correctly.

6) *Repeats.* When most of a transmission is lost, a call should be followed by correct abbreviations to ask for repeats. When a few words on the end of a transmission are lost, the *last word received correctly* is given after ?AA, meaning "all after." When a few words on the beginning of a transmission are lost, ?AB for "all before" a stated word should be used. The quickest way to ask for a fill in the middle of a transmission is to send the last word received correctly, a question mark, then the next word received correctly. Another way is to send "?BN [word] and [word]."

Do not send words twice (QSZ) unless it is requested. Send single. Do not fall into the bad habit of sending double *without a request* from fellows you work. Don't say "QRM" or "QRN" when you mean "QRS." Don't CQ unless there is definite reason for so doing. When sending CQ, use judgment.

General Practices

When a station has receiving trouble, the operator asks the transmitting station to "QSV." The letter "R" is often used in place of a decimal point (e.g., "3R5 Mc.") or the colon in time designation (e.g., "2R30 PM"). A long dash is sent for "zero."

The law concerning superfluous signals should be noted. If you *must* test, disconnect the antenna system and use an equivalent "dummy" antenna. Send your call frequently when operating. Pick a time for adjusting the station apparatus when few stations will be bothered.

The up-to-date amateur station uses "break-in." For best results send at a medium speed. Send evenly with proper spacing. The standard-type telegraph key is best for all-round use. Regular daily practice periods, two or three periods a day, are best to acquire real familiarity and proficiency with code.

No excuse can be made for "garbled" copy. Operators should copy what is sent and refuse to acknowledge a whole transmission until every word has been received correctly. *Good operators do not guess.* "Swing" in a fist is *not* the mark of a good operator. Unusual words are sent twice, the word repeated following the transmission of "?". If not *sure*, a good operator systematically asks for a fill or repeat. Sign your call frequently, interspersed with calls, and at the end of all transmissions.

On Good Sending

Assuming that an operator has learned sending properly, and comes up with a precision "fist" — not fast, but clean, steady, making well-formed rhythmical characters and spacing beautiful to listen to — he then becomes subject to outside pressures to his own possible detriment in everyday operating. He will want to "speed it up" because the operator at the other end is going faster, and so he begins, unconsciously, to run his words together or develops a "swing."

Perhaps one of the easiest ways to get into bad habits is to do too much playing around with special keys. Too many operators spend only enough time with a straight key to acquire "passable" sending, then subject their newly-developed "fists" to the entirely different movements of bugs, side-swipers, electronic keys, or what-have-you. All too often, this results in the ruination of what may have become a very good "fist."

Think about your sending a little. Are you satisfied with it? You should not be — ever. Nobody's sending is perfect, and therefore *every* operator should continually strive for improvement. Do you ever run words together — like Q for MA, or P for AN — especially when you are in a hurry? Practically everybody does at one time or another. Do you have a "swing"? Any recognizable "swing" is a deviation from perfection. Strive to send like tape sending; copy a W1AW Bulletin and try to send it with the same spacing using a local oscillator on a subsequent transmission.

Check your spacing in characters, between characters and between words occasionally by making a recording of your fist on an inked tape recorder. This will show up your faults as nothing else will. Practice the correction of faults.

USING A BREAK-IN SYSTEM

Break-in avoids unnecessarily long calls, prevents QRM, gives more communication per hour of operating. Brief calls with frequent short pauses for reply can approach (but not equal) break-in efficiency.

A separate receiving antenna makes it possible to listen to most stations while the transmitting tubes are heated. It is only necessary with break-in to pause just a moment occasionally when the key is up (or to cut the carrier momentarily and pause in a 'phone conversation) to listen for the other station. The click when the carrier is cut off is as effective as the word "break."

C.w. telegraph break-in is usually simple to arrange. With break-in, ideas and messages to be transmitted can be pulled right through the holes in the QRM. Snappy, effective, efficient, enjoyable amateur work really requires but a simple switching arrangement in your station to cut off the power and switch 'phones from monitor to receiver.

In calling, the transmitting operator sends the letters "BK" at frequent intervals during his call so that stations hearing the call may know that break-in is in use and take advantage of the fact. *He pauses at intervals* during his call, to listen for a moment for a reply. If the station being called does not answer, the call can be continued.

A tap of the key, and the man on the receiving end can interrupt (if a word is missed) since the receiver is monitoring, awaiting just such directions constantly. It is not necessary that *you* have perfect facilities to take advantage of break-in when the stations you work are break-in-equipped. After any invitation to *break* is given (and at each pause) tap your key — and contact can start *immediately*.

VOICE OPERATING

The use of proper procedure to get best results is just as important as in using code. In telegraphy words must be spelled out letter by letter. It is therefore but natural that abbreviations and shortcuts should have come into widespread use. In voice work, however, abbreviations are not necessary, and should have less importance in our operating procedure.

The letter "K" has been agreed to in telegraphic practice so that the operator will not have to pound out the separate letters that spell the words "go ahead." The voice operator can *say* the words "go ahead" or "over," or "come in please."

One laughs on c.w. by spelling out HI. On 'phone *use* a laugh when one is called for. Be natural as you would with your family and friends.

The matter of reporting *readability* and *strength* is as important to 'phone operators as to those using code. With telegraph nomenclature, it is necessary to spell out words to describe signals or use the abbreviated signal reporting system (RST . . . see Chapter Twenty-Four). Using voice, we have the ability to "say it with words." "Readability four, Strength eight" is the best way to give a quantitative report. Reporting can be done so much more meaningfully with ordinary words: "You are weak but you are in the clear and I can understand you, so go ahead," or "Your signal is strong but you are buried under local interference." Why not say it with words?

Voice Equivalents to Code Procedure

Voice	*Code*	*Meaning*
Go ahead; over	K	Self-explanatory
Wait; stand by	$\overline{AS}$, QRX	Self-explanatory
Okay	R	Receipt for a correctly-transcribed message or for "solid" transmission with no missing portions

'Phone-Operating Practice

Efficient voice communication, like good c.w. communication, demands good operating. Adherence to certain points "on getting results" will go a long way toward improving our 'phone-band operating conditions.

Voice-Operating Hints

1) Listen before calling.

2) Make short calls with breaks to listen. Avoid long CQs; do not answer any.

3) Use push-to-talk. Give essential data concisely in first transmission.

4) Make reports honest. Use definitions of strength and readability for reference. Make your reports informative and useful. Honest reports and *full* word description of signals save amateur operators from FCC trouble.

5) Limit transmission length. Two minutes or less will convey much information. When three or more stations converse in round tables, brevity is essential.

6) Display sportsmanship and courtesy. Bands are congested . . . make transmissions meaningful . . . give others a break.

7) Check transmitter adjustment . . . avoid AM overmodulation and splatter. Do not radiate when moving VFO frequency or checking NFM swing. Use receiver b.f.o. to check stability of signal. Complete testing before busy hours!

Use push-to-talk technique. Where possible arrange on-off switches or controls for fast back-and-forth exchanges that emulate the practicality of the wire telephone. This will help reduce the length of transmissions and keep brother amateurs from calling you a "monologuist" — a guy who likes to hear himself talk!

Listen with care. Keep noise and "backgrounds" out of your operating room to facilitate good listening. It is natural to answer the strongest signal, but take time to listen and give some consideration to the *best* signals, regardless of strength. Every amateur cannot run a kilowatt, but there is no reason why every amateur cannot have a signal of good quality, and utilize uniform operating practices to aid in the understandability and ease of his own communications.

Interpose your call regularly and at frequent intervals. Three short calls are better than one long one. In calling CQ. one's call should certainly appear at least once for every five or six CQs. Calls with frequent breaks to listen will save time and be most productive of results. In identifying, always transmit your *own* call *last*. *Don't* say "This is W1ABC standing by for W2DEF"; say "W2DEF, this is W1ABC, over." FCC regulations *require* that the call of the transmitting station be sent *last*.

Include country prefix before call. It is not correct to say "9RRX this is 1BDI." Correct and legal use is "W9RRX this is W1BDI." FCC regulations require proper use of calls; stations have been cited for failure to comply with this requirement.

Monitor your own frequency. This helps in timing calls and transmissions. Send when there is a chance of being copied successfully — not when you are merely "more QRM." Timing transmissions is an art to cultivate.

Keep modulation constant. By turning the gain "wide open" you are subjecting anyone listening to the diversion of whatever noises are present in or near your operating room, to say nothing of the possibility of feed-back, echo due to poor acoustics and modulation excesses due to sudden loud noises. Speak near the microphone, and don't let your gaze wander all over the station causing sharply-varying input to your speech amplifier; at the same time, keep far enough from the microphone so your signal is not modulated by your breathing. Change distance or gain only as necessary to insure uniform transmitter performance without overmodulation, splatter or distortion.

Make connected thoughts and phrases. Don't mix disconnected subjects. Ask questions consistently. Pause and get answers.

Have a pad of paper handy. It is convenient and desirable to jot down questions as they come in the course of discussion in order not to miss any. It will help you to make intelligent to-the-point replies.

Steer clear of inanities and soap-opera stuff. Our amateur radio and also our personal reputation as a serious communications worker depend on us.

Avoid repetition. Don't repeat back what the other fellow has just said. Too often we hear a conversation like this: "Okay on your new antenna there, okay on the trouble you're having with your receiver, okay on the company who just came in with some ice cream, okay . . . [etc.]." Just *say* you received everything OK. Don't try to prove it.

Use phonetics only as required. When clarifying genuinely doubtful expressions and in getting your call identified positively we suggest use of the ARRL Phonetic List. Limit such use to really-necessary clarification.

The speed of radiotelephone transmission (with perfect accuracy) depends almost entirely upon the skill of the two operators involved. One must learn to speak at a rate allowing perfect understanding as well as permitting the receiving operator to copy down the message text, if that is necessary. Because of the similarity of many English speech sounds, the use of alphabetical word lists has been found necessary. All voice-operated stations should use a *standard* list as needed to identify call signals or unfamiliar expressions.

ARRL Word List for Radiotelephony

ADAM	JOHN	SUSAN
BAKER	KING	THOMAS
CHARLIE	LEWIS	UNION
DAVID	MARY	VICTOR
EDWARD	NANCY	WILLIAM
FRANK	OTTO	X-RAY
GEORGE	PETER	YOUNG
HENRY	QUEEN	ZEBRA
IDA	ROBERT	

Example: W1AW . . . W 1 ADAM WILLIAM.

Round Tables. The round table has many advantages if run properly. It clears frequencies of interference, especially if all stations involved are on the same frequency, while the enjoyment value remains the same, if not greater. By use of push-to-talk, the conversation can be kept lively and interesting, giving each station operator ample opportunity to participate without waiting overlong for his turn.

Round tables can become very unpopular if they are not conducted properly. The monologuist, off on a long spiel about nothing in particular, cannot be interrupted: *make your transmissions short and to the point.* "Butting in" is discourteous and unsportsmanlike: *don't enter a round table, or any contact between two other amateurs, unless you are invited.* It is bad enough trying to understand voice through prevailing interference without the added difficulty of poor quality: *check your transmitter adjustments frequently.* In general, follow the precepts as hereinbefore outlined for the most enjoyment in round tables as well as any other form of radiotelephone communication.

WORKING DX

Most amateurs at one time or another make "working DX" a major aim. As in every other phase of amateur work, there are right and wrong ways to go about getting best results in working foreign stations, and it is the intention of this section to outline a few of them.

The ham who has trouble raising DX stations readily may find that poor transmitter efficiency is not the reason. He may find that his sending is poor, or his calls ill-timed, or his judgment in error. When conditions are right to bring in the DX, and the receiver sensitive enough to bring in several stations from the desired locality, the way to work DX is to use the appropriate frequency and timing and *call these stations*, as against the common practice of calling "CQ DX."

The call CQ DX means slightly different things to amateurs in different bands:

a) On v.h.f., CQ DX is a general call ordinarily used only when the band is open, under favorable "skip" conditions. For v.h.f. work such a call is used for looking for new states and countries, also for distances beyond the customary "line-of-sight" range on most v.h.f. bands.

b) CQ DX on our 7-, 14- and 28-Mc. bands may be taken to mean "General call to any foreign station." The term "foreign station" usually refers to any station in a foreign continent. (*Experienced* amateurs in the U. S. A. and Canada do *not* use this call, but *answer* such calls made by foreign stations.)

c) CQ DX used on 3.5 Mc. under winter-night conditions may be used in this same manner. At other times, under average 3.5-Mc. propagation conditions, the call may be used in domestic work when looking for new states or countries in one's own continent, usually applying to stations located over 1000 miles distant from your own.

The way to work DX is not to use a CQ call at *all* (in our continent). Instead, use your best tuning skill — and listen — and listen — and *listen. You have to hear them before you can work them.* Hear the desired stations first; time your calls well. Use your utmost skill. A sensitive receiver is often more important than the power input in working foreign stations. Before you can expect to be successful in working any particular foreign country or area, you should be able to hear ten or a dozen stations from that area.

One of the most effective ways to work DX is to know the operating habits of the DX stations sought. Doing too much transmitting on the DX bands is not the way to do this. Again, *listening* is effective. Once you know the operating habits of the DX station you are after you will know when and where to call, and when to remain silent waiting your chance.

Many DX stations use the signals HM, MH, LM and ML to indicate where they are tuning for replies. The meanings of these signals are as follows:

HM — Will start to listen at *high*-frequency end of band and tune toward *middle* of band.

MH — Will start to listen in the *middle* of the band and tune toward the *high*-frequency end.

LM — Will start to listen at *low*-frequency end of band and tune toward *middle* of band.

ML — Will start to listen in the *middle* of the band and

DX OPERATING CODE
(For W/VE Amateurs)

Some amateurs interested in DX work have caused considerable confusion and QRM in their efforts to work DX stations. The points below, if observed by all W/VE amateurs, will go a long way toward making DX more enjoyable for everybody.

1. Call DX only after he calls CQ, QRZ?, signs SK, or 'phone equivalents thereof.

2. Do *not* call a DX station:
   a. On the frequency of the station he is working until you are *sure* the QSO is over. This is indicated by the ending signal $\overline{\text{SK}}$ on c.w. and any indication that the operator is listening, on 'phone.
   b. Because you hear someone else calling him.
   c. When he signs $\overline{\text{KN}}$, $\overline{\text{AR}}$, CL, or 'phone equivalents.
   d. Exactly on his frequency.
   e. After he calls a directional CQ, unless of course you are in the right direction or area.

3. Keep within frequency-band limits. Some DX stations operate outside. Perhaps they can get away with it, but you cannot.

4. Observe calling instructions of DX stations. "10U" means call ten kc. *up* from his frequency, "15D" means 15 kc. *down*, etc.

5. Give honest reports. Many foreign stations *depend* on W and VE reports for adjustment of station and equipment.

6. Keep your signal clean. Key clicks, chirps, hum or splatter give you a bad reputation and may get you a citation from FCC.

7. *Listen* for and *call* the station you want. Calling CQ DX is not the best assurance that the *rare* DX will reply.

8. When there are several W or VE stations waiting to work a DX station, avoid asking him to "listen for a friend." Let your friend take his chances with the rest. Also avoid engaging DX stations in rag-chews against their wishes.

DATE TIME	STATION CALLED	CALLED BY	HIS FREQ. OR DIAL	HIS SIGNALS RST	MY SIGNALS RST	FREQ. MC.	EMISSION TYPE	POWER INPUT WATTS	TIME OF ENDING QSO	OTHER DATA
10-20-47										
6:15 PM	WØTQD	x	3.65	589x	569x	3.5	A-1	250	6:43	Lots of tfc! Rec'd 6, sent 10.
7:20	CQ	x				7				
7:21	x	W4TWI	7.24	369	579x				7:32	Too much QRM! Gave it up.
9:32	W3UA	x				3.95	A-3	100		Guess I was snowed under.
10-21-47										
7:05 AM	VK4DY	x	14.03			14	A-1	250		Answered a W6
7:07	AC4YN	x	14.02							ND
7:09	VK2ADW	x	14.07	339	559x				7:20	Sydney, Australia First VK!!
7:31	CQ	x								No luck
7:42	W6RBQ	x	14.05	589	579				8:02	Had to QRT for breakfast. Nice chat.
8:02		— off								

KEEP AN ACCURATE AND COMPLETE STATION LOG AT ALL TIMES! F.C.C. REQUIRES IT.

A page from the official ARRL log is shown above, answering every Government requirement in respect to station records. Bound logs made up in accord with the above form can be obtained from Headquarters for a nominal sum or you can prepare your own, in which case we offer this form as a suggestion. The ARRL log has a special wire binding and lies perfectly flat on the table.

tune toward the *low*-frequency end.

Example: If the procedure will be to tune from the middle of the band to the high end, a CQ call goes: CQ DE G5BY MH K.

ARRL has recommended some operating procedures to DX stations aimed at controlling some of the thoughtless operating practices sometimes used by W/VE amateurs. A copy of these recommendations (Operating Aid No. 5) can be obtained free of charge from ARRL Headquarters.

In any band, particularly at line-of-sight frequencies, when directional antennas are used, the directional CQ such as CQ W5, CQ north, etc., is the preferable type of call. Mature amateurs agree that CQ DX is a wishful rather than a practical type of call for most stations in the North Americas looking for contacts in foreign countries. Ordinarily, it is a cause of unnecessary QRM.

Conditions in the transmission medium make all field strengths from a given region more nearly equal at a distance, irrespective of power used. In general, the higher the frequency band, the less important power considerations become.

KEEPING AN AMATEUR STATION LOG

The FCC requires every amateur to keep a complete station operating record. It may also contain records of experimental tests and adjustment data. A stenographer's notebook can be ruled with vertical lines in any form to suit the user. The Federal Communications Commission requirements are that a log be maintained that shows (1) the date and time of *each* transmission, (2) *all calls* and transmissions made (whether two-way contacts resulted or not), (3) the input power to the last stage of the transmitter, (4) the frequency band used, (5) the time of *ending* each QSO and the operator's identifying signature for responsibility for each session of operating. Messages may be written in the log or separate records kept — but record must be made for one year as required by the FCC. For the convenience of amateur station operators ARRL stocks both logbooks and message blanks, and if one uses the official log he is sure to comply fully with the Government requirements if the precautions and suggestions included in the log are followed.

Message Handling

Amateur operators in the United States and a few other countries enjoy a privilege not available to amateurs in most countries — that of handling third-party message traffic. In the early history of amateur radio in this country, some amateurs who were among the first to take advantage of this privilege formed an extensive relay organization which became known as the American Radio Relay League.

Thus, amateur message-handling has had a long and honorable history and, like most services, has gone through many periods of development and change. Those amateurs who handled traffic in 1914 would hardly recognize it the way some of us do it today, just as equipment in those days was far different from that in use now. Progress has been made and new methods have been developed in step with advancement in communication techniques of all kinds. Amateurs who handled a lot of traffic found that organized operating schedules were more effective than random relays, and as techniques advanced and messages increased in number, trunk lines were organized, spot frequencies began to be used, and there sprang into existence a number of traffic nets in which many stations operated on the same frequency to effect wider coverage in less time with fewer relays; but the old methods are still available to the amateur who handles only an occasional message.

Although message handling is as old an art as is amateur radio itself, there are many amateurs who do not know how to handle a message and have never done so. As each amateur grows older and gains experience in the amateur service, there is bound to come a time when he will be called upon to handle a written message, during a communications emergency, in casual contact with one of his many acquaintances on the air, or as a result of a request from a nonamateur friend. Regardless of the occasion, if it comes to you, you will want to rise to it! Considerable embarrassment is likely to be experienced by the amateur who finds he not only does not know the form in which the message should be prepared, but does not know what to do with the message once it has been filed or received in his station.

Traffic work need not be a complicated or time-consuming activity for the casual or occasional message-handler. Amateurs may participate in traffic work to whatever extent they wish, from an occasional message now and then to becoming a part of organized traffic systems. This chapter explains some principles so the reader may know where to find out more about the subject and may exercise the message-handling privilege to best effect as the spirit and opportunity arise.

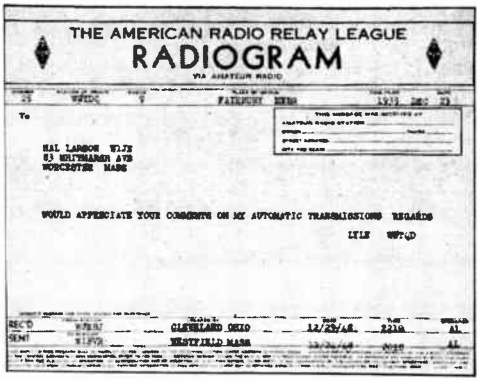
Responsibility

Amateurs who originate messages for transmission or who receive messages for relay or delivery should first consider that in doing so they are accepting the responsibility of clearing the message from their station on its way to its destination in the shortest possible time. Forty-eight hours after filing or receipt is the generally-accepted rule among traffic-handling amateurs, but it is obvious that if every amateur who relayed the message allowed it to remain in his station this long it might be a long time reaching its destination. Traffic should be relayed or delivered as quickly as possible.

Message Form

Once this responsibility is realized and accepted, handling the message becomes a matter of following generally-accepted standards of form and transmission. For this purpose, each message is divided into four parts: the preamble, the address, the text and the signature. Some of these parts themselves are subdivided. It is necessary in preparing the message for transmission and in actually transmitting it to know not only what each part is and what it is for, but to know in what *order* it should be transmitted, and to know the various procedure signals used with it when sent by c.w. If you are going to send a message, you may as well send it right.

Standardization is important! There is a great deal of room for expressing originality and individuality in amateur radio, but there are also times and places where such expression can only cause confusion and inefficiency. Recognizing the need for standardization in



Here is an example of a plain-language message in correct ARRL form, carrying the landline check.

message form and message transmitting procedures, ARRL has long since recommended such standards, and most traffic-interested amateurs have followed them. In general, these recommendations, and the various changes they have undergone from year to year, have been at the request of amateurs participating in this activity, and they are completely outlined and explained in *Operating an Amateur Radio Station*, a copy of which is available upon request or by use of the coupon at the end of Chapter Twenty-Three.

Clearing a Message

Amateurs not experienced in message handling should depend on the experienced message-handler to get a message through, if it is important; but the average amateur can enjoy operating with a message to be handled either through a local traffic net or by free-lancing. The latter may be accomplished by careful listening for an amateur station at desired points, directional CQs, use of the General Calling frequencies, or by making and keeping a schedule with another amateur for regular work between specified points. He may well aim at learning and enjoying through doing. The joy and accomplishment in thus developing one's operating skill to top perfection has a reward all its own.

The best way to clear a message is to put it into one of the many organized traffic networks, or to give it to a station who can do so. There are many amateurs who make the handling of traffic their principal operating activity, and many more still who participate in this activity to a greater or lesser extent. The result is a system of traffic nets which spreads to all corners of the United States and covers most U. S. possessions and Canada. Once a message gets into one of these nets, regardless of the net's size or coverage, it is systematically routed toward its destination in the shortest possible time.

If you decide to "take the bull by the horns" and put the message into a traffic net yourself (and more power to you if you do!), you will need to know something about how traffic nets operate, and the special Q signals and procedure they use to dispatch all traffic with a maximum of efficiency. Reference to net lists in *QST* (usually in the November and January issues) will give you the frequency and operating time of the net in your section, or other net into which your message can go. Listening for a few minutes at the time and frequency indicated should acquaint you with enough fundamentals to enable you to report into the net and indicate your traffic. From that time on you follow the instructions of the net control station, who will tell you when and to whom (and on what frequency, if different from the net frequency) to send your message. Since most nets use the special "QN" signals, it is usually very helpful to have a list of these before you (list available from ARRL Hq.).

Network Operation

About this time, you may find that you are enjoying this type of operating activity and want to know more about it, and to increase your proficiency. Many amateurs are happily "addicted" to traffic handling after only one or two brief exposures to it. Most traffic nets are at present being conducted by c.w., since this mode of communication seems to be more popular for record purposes — but this does not mean that high code speed is a necessary prerequisite to working in traffic networks. There are many nets organized specifically for the slow-speed amateur, and most of the so-called "fast" nets are usually glad to slow down to accommodate slower operators, especially those nets at state or section level.

The significant facet of net operation, however, is that code speed alone does *not* make for efficiency — sometimes quite the contrary! A high-speed operator who does not know net procedure can "foul up" a net much more completely and more quickly than can a slow operator. It is a proven fact that a bunch of high-speed operators who are not "savvy" in net operation cannot accomplish as much during a specified period as an equal number of slow operators who *know* net procedure. Don't let your code speed deter you from getting into traffic work. Given a little time, your speed will reach the point where you can compete with the best of them. Concentrate first on learning net procedure, for most traffic nowadays is handled on nets.

Team work is the theme of net operation. The net which functions most efficiently is the net in which all participants are thoroughly familiar with the procedure used, and in which operators refrain from transmitting except at the direction of the net control station, and do not occupy time with extraneous comments, even exchange of pleasantries. There is a time and place for everything. When a net is in session it should concentrate on handling traffic until all traffic is cleared. Before or after the net is the time for rag-chewing and discussion.

Some details of net operation are included in *Operating an Amateur Radio Station*, mentioned earlier, but the whole story cannot be told. There is no substitute for actual participation.

The National Traffic System

To facilitate and speed the movement of message traffic, ARRL has adopted for trial the plans urged by leading traffic men for an integrated national system by means of which originated traffic will normally reach its destination area the same day the message is originated. This system uses the local section net as a basis. Each section net sends a representative to a "regional" net (normally covering a call area) and each "regional" net sends a representative to an "area" net (normally covering a time zone). After the area net has cleared all its traffic, its members then go back to their respective regional nets, where they clear traffic to the various section net representatives. When this is done, the section representatives return to their section nets to distribute the traffic to or near its ultimate destination. By means of connecting schedules between the four area nets, traffic can flow both ways so that traffic originated on the West Coast reaches the East Coast the same night it is originated, and vice versa. In general local section nets function at 1900, regional nets at 1945, area nets at 2030 and the same or different regional and section groups meet again at 2115 and 2200 respectively. Local time is referred to in each case.

The NTS plan somewhat spreads traffic opportunity so that casual traffic may be reported into nets for efficient handling one or two nights per week, early or late; or the ardent traffic man can operate in *both* early and late groups and in between to roll up impressive totals and speed traffic reliably to its destination. Old-time traffic men who prefer a high degree of organization and teamwork have returned to the traffic game as a result of the new system. Beginners have shown more interest in becoming part of a system nationwide in scope, in which *anyone* can participate. The National Traffic System has vast and intriguing possibilities as an amateur service.

The above is but the briefest résumé of what is of necessity a rather complicated arrangement of nets and schedules. Complete details of the System and its operation are available to anyone interested. Just drop a line to ARRL Headquarters.

CHAPTER 22

Emergency Communication

One of the most important ways in which the amateur serves the public, thus making his existence a national asset, is by his preparation for and his participation in communications emergencies. Every amateur, regardless of the extent of his normal operating activities, should give some thought to the possibility of his being the only means of communication should his community be cut off from the outside world. It has happened many times, often in the most unlikely places; it has happened without warning, finding some amateurs totally unprepared; it can happen to *you.* Are you ready?

There are two principal ways in which any amateur can prepare himself for such an eventuality. One is to provide himself with equipment capable of operating on any type of emergency power (i.e., either a.c. or d.c.), and equipment which can readily be transported to the scene of disaster. Mobile equipment is especially desirable in most emergency situations.

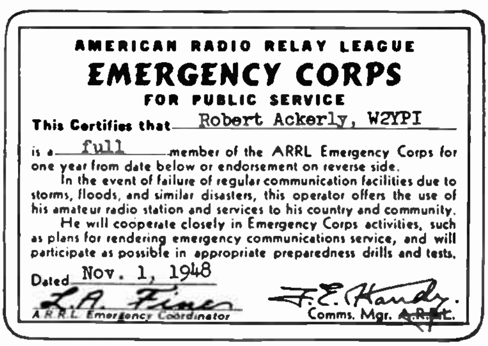
Such equipment, regardless of its elaborateness or modernness, is of little use, however, if it is not used properly and at the right times; and so another way for an amateur to prepare himself for emergencies, by no means less important than the first, is to *learn to operate efficiently.* There are many amateurs who feel that they know how to operate efficiently who find themselves considerably handicapped at the crucial time by not knowing proper procedure, by being unable due to years of casual amateur operation to adapt themselves to snappy, abbreviated transmissions, and by being unfamiliar with message form and routing procedures. It is dangerous to overrate your ability in this respect; it is far better to assume that you have much to learn.

In general it can be said that there is more emergency equipment available than there are operators who know properly how to operate during emergency conditions, for such conditions require clipped, terse procedure with complete break-in on c.w. and fast push-to-talk on 'phone. The casual rag-chewing aspect of amateur radio, however enjoyable and worth while in its place, must be forgotten at such times in favor of the business at hand.

There is only one way to gain experience in this type of operation, and that is by practicing it. During an emergency is no time for practice; it should be done beforehand, as often as possible, on a regular basis.

This leads up to the necessity for emergency organization and preparedness. ARRL has long recognized this necessity and has provided for it. The Section Communications Manager (whose address appears on page 6 of any recent issue of *QST*) is empowered to appoint certain qualified amateurs in his section for the purpose of coördinating emergency communication organization and preparedness in specified areas or communities. This appointee is known as an Emergency Coördinator for the city or town. One is specified for each community. For coördination and promotion at section level a Section Emergency Coördinator arranges for and recommends the appointments of various Emergency Coördinators at activity points throughout the section. Emergency Coördinators organize amateurs in their communities according to local needs for emergency communication facilities.

The community amateurs taking part in the local organization are members of the ARRL Emergency Corps (AEC). *All* amateurs are invited to register in the AEC, whether they are able to play an active part in their local organization or only a supporting rôle. Application blanks are available from your Emergency Coördinator, from your Section Emergency


AMERICAN RADIO RELAY LEAGUE
EMERGENCY CORPS
FOR PUBLIC SERVICE
This Certifies that Robert Ackerly, W2YPI
is a full member of the ARRL Emergency Corps for one year from date below or endorsement on reverse side.
In the event of failure of regular communication facilities due to storms, floods, and similar disasters, this operator offers the use of his amateur radio station and services to his country and community.
He will coöperate closely in Emergency Corps activities, such as plans for rendering emergency communications service, and will participate as possible in appropriate preparedness drills and tests.
Dated Nov. 1, 1948
A.R.R.L. Emergency Coördinator
Comms. Mgr. A.R.R.L.

EMERGENCY CORPS MEMBERSHIP CARD
Have You Got Yours?

Coördinator, from your Section Communications Manager or direct from ARRL Headquarters. In the event that inquiry reveals no Emergency Coördinator appointed for your community, your SCM would welcome a recommendation either from yourself or from a radio club of which you are a member. By holding an amateur operator license, you have the responsibility both to your community and to amateur radio to uphold the traditions of the service.

Among the League's publications is a booklet entitled *Emergency Communications*. This booklet, while small in size, contains a wealth of information on AEC organization and functions and is invaluable to any amateur participating in emergency work. It is free to AEC members and should be in every amateur's shack. Drop a line to the ARRL Communications Department if you want a copy, or use the coupon at the end of Chapter Twenty-Three.

Before Emergency

PREPARE yourself by providing a transmitter-receiver set-up together with an emergency power source upon which you can depend.

TEST both the dependability of your emergency equipment and your own operating ability in the annual ARRL Field Day and the several other on-the-air contests which take place annually.

REGISTER your facilities and your availability with your local ARRL Emergency Coördinator. If your community has no EC, contact your local civic and relief agencies and explain to them what the Amateur Service offers the community in time of disaster.

In Emergency

LISTEN before you transmit. Never violate this principle.

REPORT at once to your Emergency Coördinator so that he will have up-to-the-minute data on the facilities available to him. Work with local civic and relief agencies as the EC suggests, offer these agencies your services directly in the absence of an EC.

RESTRICT all on-the-air work in accordance with FCC regulations, Sec. 12.156, as soon as FCC has "declared" a state of communications emergency.

QRRR is the official ARRL "land SOS," a distress call for emergency only. It is for use *only* by a station seeking assistance.

RESPECT the fact that the success of the amateur effort in emergency depends largely on circuit discipline. The key station *in the emergency zone* should be the supreme authority for priority and traffic routing.

CO-OPERATE with those we serve. Be ready to help, but stay off the air unless there is a specific job to be done that you can handle more efficiently than any other station.

COPY all bulletins from W1AW. During time of emergency special bulletins will keep you posted on the latest developments.

After Emergency

REPORT to ARRL Headquarters as soon as possible and as fully as possible so that the Amateur Service can receive full credit. Amateur Radio has won glowing public tribute in over 75 major disasters since 1919. Maintain this record.

CHAPTER 23

ARRL Operating Organization

Amateur operation must have point and constructive purpose to win public respect. Each individual amateur is the ambassador of the entire fraternity in his public relations and attitude toward his hobby. ARRL field organization adds point and purpose to amateur operating.

The Communications Department of the League is concerned with the practical operation of stations in all branches of amateur activity. Appointments or awards are available for rag-chewer, traffic enthusiast, 'phone operator, DX man and experimenter.

There are seventy-two ARRL Sections in the League's field organization, which embraces the United States, Canada and certain other territory. Operating affairs in each Section are supervised by a Section Communications Manager elected by members in that section for a two-year term of office. Organization appointments are made by the section managers. The election of officials is covered in detail in the League's Constitution and By-Laws. Section communications managers' addresses for all sections are given in full in each issue of *QST*. SCMs welcome monthly activity reports from all amateur stations in their jurisdiction. Full information on appointments may be obtained from SCMs and is also contained in *Operating an Amateur Radio Station.*

Whether your activity embraces 'phone or telegraphy, or both, there is a place for you in League organization.

LEADERSHIP POSTS

To advance each type of station work and group interest in amateur radio, and to develop practical communications plans with the greatest success, appointments of leaders and organizers in particular single-interest fields are made by SCMs. Each leadership post is important. Each provides activities and assistance for appointee groups and individual members along the lines of natural interest. While some posts further the general ability of amateurs to communicate efficiently at all times, by pointing activity toward networks and round tables, others are aimed specifically at establishment of provisions for organizing the amateur service as a stand-by communications group to serve the public in disaster or emergency of any sort. The SCM appoints the following in accordance with section needs and individual qualifications:

PAM	'Phone Activities Manager. Organizes activities for OPSs and voice operators in his section.
RM	Route Manager. Coördinates traffic activities.
SEC	Section Emergency Coördinator. Promotes and administers section emergency radio organization.
EC	Emergency Coördinator. Organizes amateurs of a community or other area for emergency radio service; liaison with officials and agencies served; also with other local communication facilities.

STATION APPOINTMENTS

ARRL's field organization has a place for every active amateur who has a station. The Communications Department organization


The American Radio Relay League, Inc.
COMMUNICATIONS DEPARTMENT
Official Relay Station

exists to increase individual enjoyment in amateur radio work, and we extend a cordial invitation to every amateur to participate fully in the activities and to apply to the SCM for one of the following station appointments:

OPS	Official 'Phone Station. Voice operating, example in setting operating standards, activities on voice.
ORS	Official Relay Station. Traffic service, operates nets and trunk lines.
OBS	Official Bulletin Station. Transmits ARRL and FCC bulletin information to amateurs.
OES	Official Experimental Station. Experimental operating, collects reports v.h.f.-u.h.f.-s.h.f. propagation data, may engage in facsimile, TT, TV, etc., experiments.
OO	Official Observer. Sends coöperative notices to amateurs to assist in frequency observance, insures high-quality signals, and prevents FCC trouble.

Emblem Colors

Members wear the emblem with black-enamel background. A red background for an emblem will indicate that the wearer is SCM. SECs, ECs, RMs, PAMs may wear the emblem with green background. Observers and all station appointees are entitled to wear emblems with blue background.

SECTION NETS AND TRUNK LINES

Amateurs can add much experience and pleasure to their own amateur lives, and substance and accomplishment to the credit of all of amateur radio, when organized into effective interconnection of cities and towns.

The successful operation of a net depends a lot on the Net Control Station. This station should be chosen carefully and be one that will not hesitate to enforce each and every net rule and set the example in his own operation.

A progressive net grows, obtaining new members both directly and through other net members. Bulletins may be issued at intervals to keep in direct contact with the members regarding general net activity, to keep tab on net procedure and make suggestions for improvement, to keep track of active members and weed out inactive ones.

Official Relay Stations at key points are organized in trunk-line formation, covering fourteen east-west and north-south routes, connecting with numerous section and local networks and feeder systems for the purpose of efficient dispatch of traffic. Speedy and reliable work is carried on, the operation entirely on separate spot frequencies in the 3.5-Mc. amateur band. A station must hold ORS appointment to be considered for a trunk-line post.

Radio Club Affiliation

ARRL is pleased to grant affiliation to any amateur society having (1) 51% of the voting club membership made up of licensed United States or Canadian amateurs, and (2) 51% of its licensed amateurs also members of ARRL. Where a society has common aims and wishes to add strength to that of other club groups to strengthen amateur radio by affiliation with the national amateur organization, a request addressed to the Communications Manager will bring the necessary forms and information to initiate the application for affiliation. Such clubs receive field-organization bulletins and special information at intervals for posting on club bulletin boards or for relay to their memberships. A travel plan providing communications, technical and secretarial contact from the Headquarters is worked out seasonally to give maximum benefits to as many as possible of the more than four hundred affiliated radio clubs. Papers on club work, suggestions for organizing, for constitutions, for radio courses of study, etc., are available on request.

Club Training Aids

One section of the ARRL Communications Department handles the Training Aids Program. This program is a service to ARRL affiliated clubs. Material is supplied for club programs aimed at education, training and entertainment of club members, to make your club meetings more interesting and consequently better attended.

Training Aids include such items as motion-picture films, film strips, slides, recordings, and lecture outlines. Also, code-proficiency training equipment such as recorders, tape transmitters and tapes will be loaned when such items are available.

All Training Aids materials are loaned free (except for shipping charges) to ARRL affiliated clubs. Numerous groups use this ARRL service to good advantage. If your club is affiliated but has not yet taken advantage of this service, you are missing a good chance to add the available features to your meeting programs and general club activities. Watch club bulletins and *QST* or write the ARRL Communications Department for full details.

W1AW

The Maxim Memorial Station, W1AW, is dedicated to fraternity and service. Operated by the League headquarters, W1AW is located about four miles south of the Headquarters offices on a seven-acre site. The station is on the air daily, except holidays, and available time is divided between different bands and modes. Telegraph and 'phone transmitters are provided for all bands from 1.8 to 144 Mc. The normal frequencies in each band for c.w. and voice transmissions are as follows: 1887, 3555, 3950, 7215, 14,100, 14,280, 28,060, 29,000, 52,000 and 146,000 kc. Operating-visiting hours and the station schedule are listed every other month in *QST*.

All amateurs are invited to visit W1AW, as well as to work the station from their own shacks. The station was established to be a living memorial to Hiram Percy Maxim and to carry on the work and traditions of the amateur fraternity.

OPERATING ACTIVITIES

Within the ARRL field organization there are several special activities. The first Saturday night each month is set aside for all ARRL officials, officers and directors to get together over the air from their own stations. This activity is known to the gang as LO-NITE. For all appointees, quarterly tests called CD parties are scheduled to develop operating ability and a spirit of fraternalism.

In addition to these special activities for appointees and members, ARRL sponsors various other activities open to all amateurs. The DX-minded amateur may participate in the Annual ARRL International DX Competition during February and March. This popular contest may bring you the thrill of working new countries. Then there is the ever-popular Sweepstakes in November. Of domestic scope, the SS affords the opportunity to work new states for that WAS award. For the 28-Mc. gang there is the Ten-Meter WAS Contest held each January. The interests of v.h.f. enthusiasts are also provided for in special activities planned by ARRL.

As in all our operating, the idea of having a good time is combined in the Annual Field Day, with the more serious thought of preparing ourselves to render public service in times of emergency. A premium is placed on the use of equipment without connection to commercial power sources. Clubs and individual groups always have a good time in the "FD," learn much about the requirements for knockabout conditions afield.

ARRL contest activities are diversified to appeal to all operating interests, and will be found announced in detail in issues of *QST* preceding the different events.

● AWARDS

The League-sponsored operating activities heretofore mentioned have useful objectives and provide much enjoyment for members of the fraternity. Achievement in amateur radio is recognized by various certificates offered through the League and detailed below.

WAS Award

WAS means "Worked All States." This award is available regardless of affiliation or nonaffiliation with any organization. Here are the few simple rules to follow in applying for a WAS Certificate:

1) Two-way communications must be established on the amateur bands with all forty-eight United States; any and all amateur bands may be used. A card from the District of Columbia may be submitted in lieu of one from Maryland.

2) Contacts with all forty-eight states must be made from the same location. Within a given community one location may be defined as from places no two of which are more than 25 miles apart.

3) Contacts may be made over any period of years, and may have been made any number of years ago, provided only that all contacts are from the same location.

4) Forty-eight QSL cards, or other written communications from stations worked confirming the necessary two-way contacts, must be submitted to ARRL headquarters.

5) Sufficient postage must be sent with the confirmations to finance their return. No correspondence will be returned unless sufficient postage is furnished.

6) The WAS award is available to all amateurs.

7) Address all applications and confirmations to the Communications Department, ARRL, 38 La Salle Road, West Hartford, Conn.

DX Century Club Award

Here are the rules under which the DX Century Club Award will be issued to amateurs who have worked and confirmed contact with 100 countries in the postwar period. If you worked fewer than 100 countries before the war and have since worked and confirmed a sufficient number to make the 100 mark, the DXCC is still available to you under the rules detailed on page 74 of June, 1946, *QST*.

1) The Century Club Award Certificate for confirmed contacts with 100 or more countries is available to all amateurs everywhere in the world.

2) Confirmations must be submitted direct to ARRL headquarters for all countries claimed. Claims for a total of 100 countries must be included with first application. Confirmation from foreign contest logs may be requested in the case of the ARRL International DX Competition only, subject to the following conditions:

a) Sufficient confirmations of other types must be submitted so that these, plus the DX Contest confirmations, will total 100. In every case, Contest confirmations must not be requested for any countries from which the applicant has regular confirmations. That is, contest confirmations will be granted only in the case of countries from which applicants have no regular confirmations.

b) Look up the contest results as published in *QST* to see if your man is listed in the foreign scores. If he isn't, he did not send in a log and no confirmation is possible.

c) Give year of contest, date and time of QSO.

d) In future DX Contests, do not request confirmations until after the final results have been published, usually in one of the early fall issues. Requests before this time must be ignored.

3) The ARRL Countries List, printed periodically in *QST*, will be used in determining what constitutes a "country." The Miscellaneous Data chapter of this *Handbook* contains the Postwar Countries List.

4) Confirmations must be accompanied by a list of claimed countries and stations to aid in checking and for future reference.

5) Confirmations from additional countries may be submitted for credit each time ten additional confirmations are available. Endorsements for affixing to certificates and showing the new confirmed total (110, 120, 130, etc.) will be awarded as additional credits are granted. ARRL DX Competition logs from foreign stations may be utilized for these endorsements, subject to conditions stated under (2).

6) All contacts must be made with amateur stations working in the authorized amateur bands or with other stations licensed to work amateurs.

7) In cases of countries where amateurs are licensed in the normal manner, credit may be claimed only for stations using regular government-assigned call letters. No credit may be claimed for contacts with stations in any countries in which amateurs have been temporarily closed down by special government edict where amateur licenses were formerly issued in the normal manner.

8) All stations contacted must be "land stations" . . . contacts with ships, anchored or otherwise, and aircraft, cannot be counted.

9) All stations must be contacted from the same call

area, where such areas exist, or from the same country in cases where there are no call areas. One exception is allowed to this rule: where a station is moved from one call area to another, or from one country to another, all contacts must be made from within a radius of 150 miles of the initial location.

10) Contacts may be made over any period of years from November 15, 1945, provided only that all contacts be made under the provisions of Rule 9, and by the same station licensee; contacts may have been made under different call letters in the same area (or country), if the licensee for all was the same.

11) All confirmations must be submitted exactly as received from the stations worked. Any altered or forged confirmations submitted for CC credit will result in disqualification of the applicant. The eligibility of any DXCC applicant who was ever barred from DXCC to reapply, and the conditions for such application, shall be determined by the Awards Committee. Any holder of the Century Club Award submitting forged or altered confirmations must forfeit his right to be considered for further endorsements.

12) OPERATING ETHICS: Fair play and good sportsmanship in operating are required of all amateurs working toward the DX Century Club Award. In the event of specific objections relative to continued poor operating ethics an individual may be disqualified from the DXCC by action of the ARRL Awards Committee.

13) Sufficient postage for the return of confirmations must be forwarded with the application. In order to insure the safe return of large batches of confirmations, it is suggested that enough postage be sent to make possible their return by first-class mail, registered.

14) Decisions of the ARRL Awards Committee regarding interpretation of the rules as here printed or later amended shall be final.

15) Address all applications and confirmations to the Communications Department, ARRL, 38 La Salle Road, West Hartford 7, Conn.

WAC Award

The International Amateur Radio Union issues WAC (Worked All Continents) certificates to all members of member-societies who submit proof of two-way communication with at least one station on each continent. Foreign amateurs submit their proof direct to member-societies of the IARU. Others may make application to ARRL, headquarters society of the Union. A c.w. and a telephony certificate are available. Also, special endorsement will be placed on certificates upon receipt of request accompanied by proof of having worked all continents on 50 Mc.

Code Proficiency Award

Many hams can follow the general idea of a contact "by ear" but when pressed to "write it down" they "muff" the copy. The Code Proficiency Award invites every amateur to prove himself as a proficient operator, and sets up a system of awards for step-by-step gains in copying proficiency. It enables every amateur to check his code proficiency, to better that proficiency, and to receive a certification of his receiving speed.

This program is a whale of a lot of fun. The League will give a certificate to any licensed radio amateur who demonstrates that he can copy perfectly, for at least one minute, plain-language Continental code at 15, 20, 25, 30 or 35 words per minute, as transmitted during special monthly transmissions from W1AW, or from W6OWP, WØTQD and others mentioned in *QST*.


The American Radio Relay League, Inc.
Certificate of Proficiency in Reception of the Continental Code

As part of the ARRL Code Proficiency program, W1AW transmits plain-language practice material each evening, Monday through Friday, at speeds from 9 to 35 w.p.m. All amateurs are invited to use these transmissions to increase their code-copying ability. Non-amateurs are invited to utilize the lower speeds, 9, 12 and 15 w.p.m., which are transmitted for the benefit of persons studying the code in preparation for the amateur license examination. Refer to any issue of *QST* for details of the practice schedule.

Rag Chewers Club

The Rag Chewers Club is designed to encourage friendly contacts and discourage the "hello-good-by" type of QSO. Its purpose is to bond together operators interested in honest-to-goodness rag-chewing over the air. Membership certificates are available.

How To Get in: (1) Chew the rag with a member of the club for at least a solid half hour. This does not mean a half hour spent in trying to get a message over through bad QRM or QRN, but a solid half hour of conversation or message handling. (2) Report the conversation by card to The Rag Chewers Club, ARRL, Communications Department, West Hartford, Conn., and ask the member station you talk with to do the same. When both reports are received you will be sent a membership certificate entitling you to all the privileges of a Rag Chewer.

How To Stay in: (1) Be a conversationalist on the air instead of one of those tongue-tied infants who don't know any words except "cuagn" or "cul," or "QRU" or "nil." Talk to the fellows you work with and get to know them. (2) Operate your station in accordance with the radio laws and ARRL practice. (3) Observe rules of courtesy on the air. (4) Sign "RCC" after each call so that others may know you can talk as well as call.

A-1 Operator Club

The A-1 Operator Club should include in its ranks every good operator. To become a member, one must be nominated by at least two operators who already belong. General keying or voice technique, procedure, copying ability, judgment and courtesy all count in rating candidates under the club rules detailed at length in *Operating an Amateur Radio Station*. Aim to make yourself a fine operator, and one of these days you will be pleasantly surprised by an invitation to belong to the A-1 Operator Club, which carries a worth-while certificate in its own right.

Brass Pounders League

Every individual reporting more than a specified minimum in official monthly traffic totals is given an honor place in the *QST* listing known as the Brass Pounders League and a certificate to recognize his performance.

The value to amateurs in operator training, and the utility of amateur message handling to the members of the fraternity itself as well as to the general public, make message-handling work of prime importance to the fraternity. Fun, enjoyment, and the feeling of having done something really worth while for one's fellows is accentuated by pride in message files, records, and letters from those served.

Old Timers Club

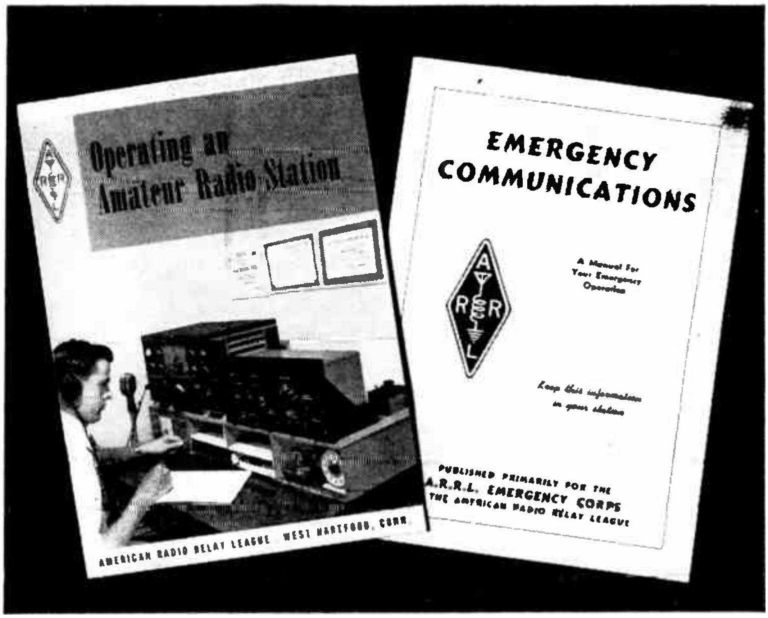
The Old Timers Club is open to anyone who holds an amateur call at the present time, and who held an amateur license (operator or station) 20-or-more years ago. Lapses in activity during the intervening years are permitted.

If you can qualify as an "Old Timer," send us a brief chronology of your ham career, being sure to indicate the date of your first amateur license, and your present call. If the evidence submitted proves you eligible for the OTC, you will be added to the roster and will receive a membership certificate.

● INVITATION

Amateur radio is capable of giving enjoyment, self-training, social and organization benefits in proportion to what the individual amateur puts into his hobby. All amateurs are invited to become ARRL members, to work toward awards, and to accept the challenge and invitation offered in field-organization appointments. Drop a line for the booklet *Operating an Amateur Radio Station*, which has detailed information on the field-organization appointments and awards. Accept today the invitation to take full part in all ARRL activities and organization work.

SEE NEXT PAGE →



▶ Operating an Amateur Radio Station covers the details of practical amateur operating. In it you will find information on Operating Practices, Emergency Communication, ARRL Operating Activities and Awards, the ARRL Field Organization, Handling Messages, Network Organization, "Q" Signals and Abbreviations used in amateur operating, important extracts from the FCC Regulations, and other helpful material. It's a handy reference that will serve to answer many of the questions concerning operating that arise during your activities on the air.

▶ If you as a licensed amateur should ever find yourself in a position to serve during an emergency, there are a lot of things you will wish you had known beforehand. You will do the best you can, and those you serve will sing your praises — but you yourself will realize that had you been better prepared you could have done *more* and done it more *effectively*. The booklet *Emergency Communications* would have told you all you needed to know. You should have had it, studied it, and followed up its advices. Don't wait until the emergency is upon you to wonder what you should do and how you should do it. Get a copy of *Emergency Communications* and make your preparations *now!*

The two publications described above may be obtained without charge by any *Handbook* reader. Either or both will be sent upon request.

AMERICAN RADIO RELAY LEAGUE
38 La Salle Road
West Hartford 7, Connecticut, U. S. A.

Please send me, without charge, the following:

☐ **OPERATING AN AMATEUR RADIO STATION**
☐ **EMERGENCY COMMUNICATIONS**

Name...
(Please Print)

Address..

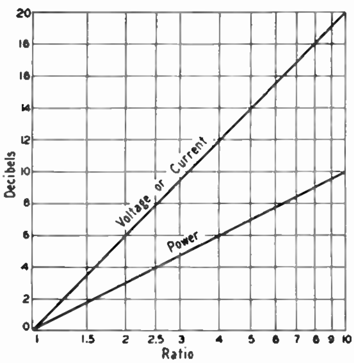
....

CHAPTER 24

Miscellaneous Data

THE DECIBEL

In most radio communication the received signal is converted into sound. This being the case, it is useful to appraise signal strengths in terms of relative loudness as registered by the ear. A peculiarity of the ear is that an increase or decrease in loudness is responsive to the *ratio* of the amounts of power involved, and is practically independent of absolute *value* of the power. For example, if a person estimates that the signal is "twice as loud" when the transmitter power is increased from 10 watts to 40 watts, he will also estimate that a 400-watt signal is twice as loud as a 100-watt signal. In other words, the ear has a *logarithmic* response.



This fact is the basis for the use of the relative-power unit called the **decibel.** A change of one decibel (abbreviated **db.**) in the power level is just detectable as a change in loudness under ideal conditions. The power ratio and decibels are related by the following formula:

$$Db. = 10 \log \frac{P_2}{P_1}$$

Common logarithms (base 10) are used.

Note that the decibel is based on *power* ratios. Voltage or current ratios can be used, but *only when the impedance is the same for both values of voltage, or current.* The gain of an amplifier cannot be expressed correctly in db. if it is based on the ratio of the output voltage to the input voltage unless both voltages are measured across the same value of impedance. When the impedance at both points of measurement is the same, the following formula may be used for voltage or current ratios:

$$Db. = 20 \log \frac{V_2}{V_1} \text{ or } 20 \log \frac{I_2}{I_1}$$

The two formulas are shown graphically in the accompanying chart for ratios from 1 to 10. Gains (increases) expressed in decibels may be added arithmetically; losses (decreases) may be subtracted. A power decrease is indicated by prefixing the decibel figure with a minus sign. Thus +6 db. means that the power has been multiplied by 4, while −6 db. means that the power has been divided by 4. The chart may be used for other ratios by adding (or subtracting, if a loss) 10 db. each time the ratio scale is multiplied by 10, for power ratios; or by adding (or subtracting) 20 db. each time the scale is multiplied by 10 for voltage or current ratios.

Example: The power input to a transmitter is increased from 75 to 600 watts. Assuming that the efficiency is the same in both cases, the ratio of the new output power to the old is 600/75 = 8. From the chart, the signal will be increased 9 db. Note that increasing the power to 750 watts, a ratio of 10, would increase the signal to 10 db., a barely perceptible increase over 600 watts.

Example: A speech amplifier has an output of 10 watts when excited by 0.02 volt from a crystal microphone. The nominal impedance of the microphone is 50,000 ohms. In a 50,000-ohm load, the voltage developed by the 10 watts would be

$$E = \sqrt{PR} = \sqrt{10 \times 50{,}000} = \sqrt{500{,}000}$$
$$= 707 \text{ volts}$$

The voltage ratio of the amplifier therefore is 707/0.02 = 35,350. This is the same as 3.5 × 10,000. A voltage ratio of 10,000 (10^4) is equal to 4 × 20 = 80 db. From the chart, a voltage ratio of 3.5 = 11 db. Adding the two gives 11 + 80 = 91 db. as the gain of the amplifier.

Example: A transmission line is terminated in its characteristic impedance and operates without standing waves. The power put into the line is 150 watts, but the power measured at the output end is 100 watts. The ratio is 150/100 = 1.5. From the chart, this ratio is equal to 1.9 db. The loss in the line is therefore 1.9 db.

DECIMAL EQUIVALENTS OF FRACTIONS

1/32	.03125	17/32	.53125
1/16	.0625	9/16	.5625
3/32	.09375	19/32	.59375
1/8	.125	5/8	.625
5/32	.15625	21/32	.65625
3/16	.1875	11/16	.6875
7/32	.21875	23/32	.71875
1/4	.25	3/4	.75
9/32	.28125	25/32	.78125
5/16	.3125	13/16	.8125
11/32	.34375	27/32	.84375
3/8	.375	7/8	.875
13/32	.40625	29/32	.90625
7/16	.4375	15/16	.9375
15/32	.46875	31/32	.96875
1/2	.5	1	1.0

SYMBOLS FOR ELECTRICAL QUANTITIES

Admittance	Y, y
Angular velocity $(2\pi f)$	ω
Capacitance	C
Conductance	G, g
Conductivity	γ
Current	I, i
Difference of potential	E, e
Dielectric constant	K
Dielectric flux	Ψ
Energy	W
Frequency	f
Impedance	Z, z
Inductance	L
Magnetic intensity	H
Magnetic flux	Φ
Magnetic flux density	B
Magnetomotive force	F
Mutual inductance	M
Number of conductors or turns	N
Period	T
Permeability	μ
Phase displacement	θ
Power	P, p
Quantity of electricity	Q, q
Reactance	X, x
Reactance, Capacitive	X_C
Reactance, Inductive	X_L
Reluctivity	ν
Resistance	R, r
Resistivity	ρ
Susceptance	b
Speed of rotation	n
Voltage	E, e
Work	W

PILOT-LAMP DATA

Lamp No.	*Bead Color*	*Base (Miniature)*	*Bulb Type*	*RATING*	
				Volts	*Amp.*
40	Brown	Screw	T-3¼	6–8	0.15
40A[1]	Brown	Bayonet	T-3¼	6–8	0.15
41	White	Screw	T-3¼	2.5	0.5
42	Green	Screw	T-3¼	3.2	**
43	White	Bayonet	T-3¼	2.5	0.5
44	Blue	Bayonet	T-3¼	6–8	0.25
45	*	Bayonet	T-3¼	3.2	**
46[2]	Blue	Screw	T-3¼	6–8	0.25
47[1]	Brown	Bayonet	T-3¼	6–9	0.15
48	Pink	Screw	T-3¼	2.0	0.06
49[3]	Pink	Bayonet	T-3¼	2.0	0.06
[4]	White	Screw	T-3¼	2.1	0.12
49A[3]	White	Bayonet	T-3¼	2.1	0.12
50	White	Screw	G-3½	6–8	0.2
51[2]	White	Bayonet	G-3½	6–8	0.2
—	White	Screw	G-4½	6–8	0.4
55	White	Bayonet	G-4½	6–8	0.4
292[5]	White	Screw	T-3¼	2.9	0.17
292A[5]	White	Bayonet	T-3¼	2.9	0.17
1455	Brown	Screw	G-5	18.0	0.25
1455A	Brown	Bayonet	G-5	18.0	0.25

* White in G.E. and Sylvania; green in National Union Raytheon and Tung-Sol.

** 0.35 in G.E. and Sylvania; 0.5 in National Union Raytheon and Tung-Sol.

[1] **40A** and **47** are interchangeable.

[2] Have frosted bulb.

[3] **49** and **49A** are interchangeable.

[4] Replace with No. **48**.

[5] Use in 2.5-volt sets where regular bulb burns out too frequently.

ABBREVIATIONS FOR ELECTRICAL AND RADIO TERMS

Alternating current	a.c.
Ampere (amperes)	a.
Amplitude modulation	AM
Antenna	ant.
Audio frequency	a.f.
Centimeter	cm.
Continuous waves	c.w.
Cycles per second	c.p.s.
Decibel	db.
Direct current	d.c.
Electromotive force	e.m.f.
Frequency	f.
Frequency modulation	FM
Ground	gnd.
Henry	h.
High frequency	h.f.
Intermediate frequency	i.f.
Interrupted continuous waves	i.c.w.
Kilocycles (per second)	kc.
Kilovolt	kv.
Kilowatt	kw.
Magnetomotive force	m.m.f.
Medium frequency	m.f.
Megacycles (per second)	Mc.
Megohm	MΩ
Meter	m.
Microfarad	μfd.
Microhenry	μh.
Micromicrofarad	μμfd.
Microvolt	μv.
Microvolt per meter	μv.m.
Microwatt	μw.
Milliampere	ma.
Millivolt	mv.
Milliwatt	mw.
Modulated continuous waves	m.c.w.
Ohm	Ω
Power	P
Power factor	p.f.
Radio frequency	r.f.
Ultrahigh frequency	u.h.f.
Very-high frequency	v.h.f.
Volt (volts)	v.
Watt (watts)	w.

TABLE OF DIELECTRIC CHARACTERISTICS

Dielectric material [1]	*Dielectric constant (K)*	*Power factor*					*Dielectric strength (puncture voltage)* [2]	*Volume resistivity* [3] ( )
		60 cycles	*1 kc.*	*1 Mc.*	*10 Mc.*	*100 Mc.*		
Air (normal pressure)	1.0						19.8–22.8	
AlSiMag A196	5.7–6.3	2.9		0.21	0.15		240	10^{14}
Aniline formaldehyde	3–5	1–6					400	
Asphalts	2.7–3.1		2.3				25–30	
Bakelite — *See* Phenol								
Beeswax	2.9–3.2							
Casein plastics [4]	6.1–6.4			5.2–6			165	
Castor oil	4.3–4.7			7			380	
Celluloid	4–16			5–10				
Cellulose acetate [5]	6–8	3–6	4–6	4–6	5.5		300–1000	4.5×10^{10}
Cellulose nitrate [6]	4–7			2.8–5			300–780	$2–30 \times 10^{10}$
Ceresin wax	2.5–2.6			0.12–0.21				
Cresol formaldehyde	6	10					400	
Dilectene	3.57					0.33		
Ethyl cellulose	2–2.7	0.7	1.2	1.5			1500	10^{15}
Fiber	5–7.5			4.5–5			150–180	5×10^{9}
Formica MF-66	4.6–4.9		1.5	1.1			450	
Glass:								
Cobalt	7.3			0.7				
Common window	7.6–8			1.4			200–250	
Crown	6.2–7		1	1 [3]			500	
Electrical	4–5			0.5			2000	8×10^{14}
Flint	7–10		0.45	0.4				
Nonex	4.2			0.25		0.28		
Photographic	7.5			0.8–1				
Plate	6.8–7.6			0.6–0.8				
Pyrex	4.2–4.9		0.5	0.7		0.54	335	10^{14}
Gutta percha	2.5–4.9						200–500	$5 \times 10^{14}–10^{15}$
Lucite [7]	2.5–3	7	5	1.5–3	1.9		480–500	
Melamine formaldehyde	8	16					300	
Mica	2.5–8	0.2	0.3	0.2–6	0.02			2×10^{17}
Mica (clear India)	6.4–7.5	2	2	2	2		600–1500	
Mycalex	7.4			0.18			250	10^{13}
Mycalex (British)	6			0.3			350	
Mykroy	6.5–7			0.1–0.2			630	
Nylon	3.6			2.2				
Paper	2.0–2.6						1250	
Paraffin wax (solid)	1.9–2.6			0.1–0.3			300	$10^{15}–10^{19}$
Pemque	7.21			0.2				
Phenol: [8]								
Pure	5			1			400–475	1.5×10^{12}
Asbestos base	7.5			15			90–150	
Black molded	5–5.5			3.5			400–500	
Fabric base	5–6.5			3.5–11			150–500	
Mica-filled	5–6			0.8–1			475–600	
Paper base	3.8–5.5			2.5–4			650–750	$10^{10}–10^{13}$
Yellow	5.3–5.4			0.36–0.7			500	
Polyethylene	2.3–2.4	0.02	0.02	0.02–0.05			1000	10^{17}
Polyindene	3	0.04						
Polyisobutylene	2.4–2.5	0.04–5	0.05				500	10^{16}
Polystyrene [9]	2.4–2.9(2.6)	0.02	0.018	0.02	0.02	0.02	500–2500	10^{20}
Porcelain (dry process)	6.2–7.5			0.7–15			40–100	5×10^{8}
Porcelain (wet process)	6.5–7			0.6			150	
Pressboard (untreated)	2.9–4.5						125–300	
Pressboard (oiled)	5						750	
Quartz (fused)	3.5 (3.8)	0.01	0.01	0.015–0.03	0.01	0.05	200	$10^{14}–10^{18}$
Rubber (hard) [10]	2–3.5(3)			0.5–1			450	$10^{12}–10^{15}$
Shellac	2.5–4			0.09			900	10^{16}
Steatite: [11]								
"Commercial" grade	4.9–6.5	0.02	0.2	0.2	0.4	0.5		
"Low-loss" grade	4.4	0.02	0.2	0.2	0.18	0.13	150–315	$10^{14}–10^{15}$
Titanium dioxide [12]	90–170		0.1	0.1				
Urea formaldehyde [13]	5–7	3–5	2–3	2–4	4		300–550	$10^{12}–10^{13}$
Varnished cloth [14]	2–2.5			2–3			440–550	
Vinyl resins	4			1.4–1.7			400–500	10^{14}
Vitrolex	6.4			0.3				
Wood (dry oak)	2.5–6.8(3)		3.8	4.2				
Wood (paraffined maple)	4.1						115	

[1] Most data taken at 25° C.

[2] Puncture voltage, in volts per mil. Most data apply to relatively thin sections and cannot be multiplied directly to give breakdown for thicker sections without added safety factor.

[3] In ohm-cm.

[4] Includes such products as Aladdinite, Ameroid, Galalith, Erinoid, Lactoid, etc.

[5] Includes Fibestas, Lumerith, Nixonite, Plastacele, Tenite, etc.

[6] Includes Amerith, Nitron, Nixonoid, Pyralin, etc.

[7] Methylmethacrylate resin.

[8] Phenolaldehyde products include Acrolite, Bakelite, Catalin, Celeron, Dielecto, Durez, Durite, Formica, Gemstone, Heresite, Indur, Makalot, Marblette, Micarta, Opalon, Prystal, Resinox, Synthane, Textolite, etc. Yellow bakelite is so-called "low-loss" bakelite.

[9] Includes Amphenol 912A, Distrene, Intelin IN 45, Loalin, Lustron, Quartz Q, Rezoglas, Rhodolene M, Ronilla L, Styraflex, Styron, Trolitul, Victron, etc.

[10] Also known as Ebonite.

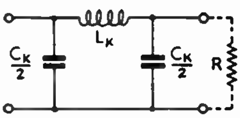
[11] Soapstone — Alberene, Alsimag, Isolantite, Lava, etc.

[12] Rutile. Used in low temperature-coefficient fixed condensers.

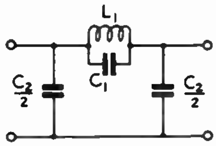
[13] Includes Aldur, Beetle, Plaskon, Pollopas, Prystal, etc.

[14] Includes Empire cloth.

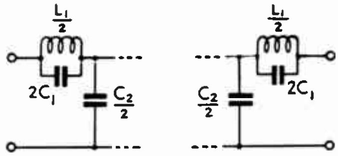
LOW-PASS FILTERS



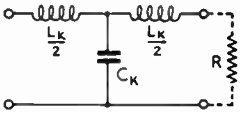
Constant-k π section



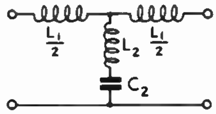
m-derived π section



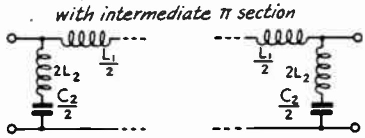
m-derived end sections for use with intermediate π section



Constant-k T section



m-derived T section



m-derived end sections for use with intermediate T section

$L_K = \frac{R}{\pi f_c}$ $C_K = \frac{1}{\pi f_c R}$

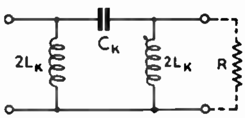
$L_1 = mL_K$ $C_1 = \frac{1-m^2}{4m} C_K$

$L_2 = \frac{1-m^2}{4m} L_K$ $C_2 = mC_K$

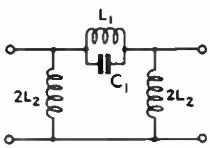
$L_1 = mL_K$ $C_1 = \frac{1-m^2}{4m} C_K$

$L_2 = \frac{1-m^2}{4m} L_K$ $C_2 = mC_K$

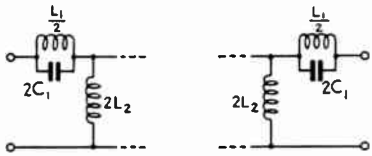
HIGH-PASS FILTERS



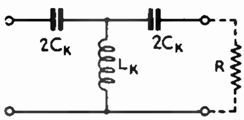
Constant-k π section



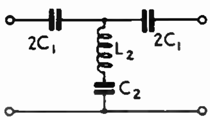
m-derived π section



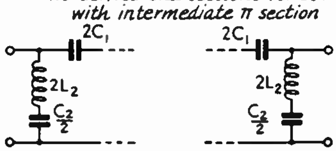
m-derived end sections for use with intermediate π section



Constant-k T section



m-derived T section



m-derived end section for use with intermediate T section

$L_K = \frac{R}{4\pi f_c}$ $C_K = \frac{1}{4\pi f_c R}$

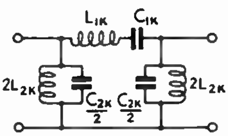
$L_1 = \frac{4m}{1-m^2} L_K$ $C_1 = \frac{C_K}{m}$

$L_2 = \frac{L_K}{m}$ $C_2 = \frac{4m}{1-m^2} C_K$

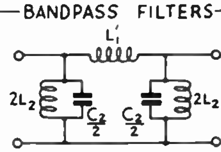
$L_1 = \frac{4m}{1-m^2} L_K$ $C_1 = \frac{C_K}{m}$

$L_2 = \frac{L_K}{m}$ $C_2 = \frac{4m}{1-m^2} C_K$

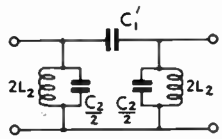
BANDPASS FILTERS



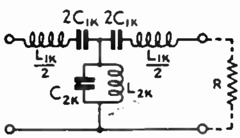
Constant-k π section



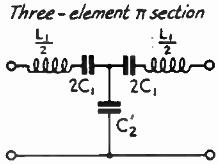
Three-element π section



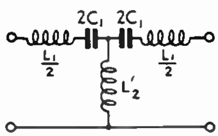
Three-element π section



Constant-k T section



Three-element T section



Three-element T section

$L_{1K} = \frac{R}{\pi(f_2-f_1)}$ $C_{1K} = \frac{f_2-f_1}{4\pi f_1 f_2 R}$

$L_{2K} = \frac{(f_2-f_1)R}{4\pi f_1 f_2}$ $C_{2K} = \frac{1}{\pi(f_2-f_1)R}$

$L_1 = L_{1K}$ $L_1' = \frac{R}{\pi(f_1+f_2)}$

$C_1 = \frac{f_2-f_1}{4\pi f_1^2 R}$ $L_2 = \frac{(f_2-f_1)R}{4\pi f_1^2}$

$C_2 = C_{2K}$ $C_2' = \frac{1}{\pi(f_1+f_2)R}$

$L_1 = \frac{f_1 R}{\pi f_2(f_2-f_1)}$ $C_1 = C_{1K}$

$C_1' = \frac{f_1+f_2}{4\pi f_1 f_2 R}$ $L_2 = L_{2K}$

$L_2' = \frac{(f_1+f_2)R}{4\pi f_1 f_2}$ $C_2 = \frac{f_1}{\pi f_2(f_2-f_1)R}$

In the above formulas *R* is in ohms, *C* in farads, *L* in henrys, and *f* in cycles per second.

FILTERS

The filter sections shown on the facing page can be used alone or, if greater attenuation and sharper cut-off are required, several sections can be connected in series. In the low- and high-pass filters, f_c represents the cut-off frequency, the highest (for the low-pass) or the lowest (for the high-pass) frequency transmitted without attenuation. In the bandpass-filter designs, f_1 is the low-frequency cut-off and f_2 the high-frequency cut-off. The units for L, C, R and f are henrys, farads, ohms and cycles, respectively.

All of the types shown are for use in an unbalanced line (one side grounded), and thus they are suitable for use in coaxial line or any other unbalanced circuit. To transform them for use in balanced lines (e.g., 300-ohm transmission line, or push-pull audio circuits), the series reactances should be equally divided between the two legs. Thus the balanced constant-k π-section low-pass filter would use two inductances of a value equal to $L_k/2$, while the balanced constant-k π-section high-pass filter would use two condensers of a value equal to $2C_k$.

If several low- (or high-) pass sections are to be used, it is advisable to use m-derived end sections on either side of a constant-k section, although an m-derived center section can be used. The factor m relates the ratio of the cut-off frequency and f_∞, a frequency of high attenuation. Where only one m-derived section is used, a value of 0.6 is generally used for m, although a deviation of 10 or 15 per cent from this value is not too serious in amateur work. For a value of $m = 0.6$, f will be $1.25f_c$ for the low-pass filter and $0.8f_c$ for the high-pass filter. Other values can be found from

$$m = \sqrt{1 - \left(\frac{f_c}{f_\infty}\right)^2}$$ for the low-pass filter and

$$m = \sqrt{1 - \left(\frac{f_\infty}{f_c}\right)^2}$$ for the high-pass filter.

The filters shown should be terminated in a resistance $= R$, and there should be little or no reactive component in the termination.

Simple audio filters can be made with powdered-iron-core chokes and paper condensers. Sharper cut-off characteristics will be obtained with more sections. The values of the components can vary by ±5% with little or no reduction in performance. The more sections there are to a filter the greater is the need for accuracy in the values of the components. High-performance audio filters can be built with only two sections by winding the inductances on toroidial powdered-iron forms — it generally takes three sections to obtain the same results when using other inductances.

Sideband filters are usually designed to operate in the range 10 to 20 kc. Their attenuation requirements are such that usually at least a five-section filter is required. The coils should be as high-Q as possible, and mica condensers are the most suitable capacitors.

Low-pass and high-pass filters for harmonic suppression and receiver-overload prevention in the television frequencies range are usually made with self-supporting coils and mica or ceramic condensers, depending upon the power requirements.

In any filter, there should be no magnetic or capacity coupling between sections of the filter unless the design specifically calls for it. This requirement makes it necessary to shield the coils from each other in some applications, or to mount them at right angles to each other.

Further information on filter design can be found in the following articles:

Bennett, "Audio Filters for Eliminating QRM," *QST*, July, 1949.

Berry, "Filter Design for the Single-Sideband Transmitter," *QST*, June, 1949.

Buchheim, "Low-Pass Audio Filters," *QST*, July, 1948.

Grammer, "Pointers on Harmonic Reduction," *QST*, April, 1949; "High-Pass Filters for TVI Reduction," *QST*, May, 1949.

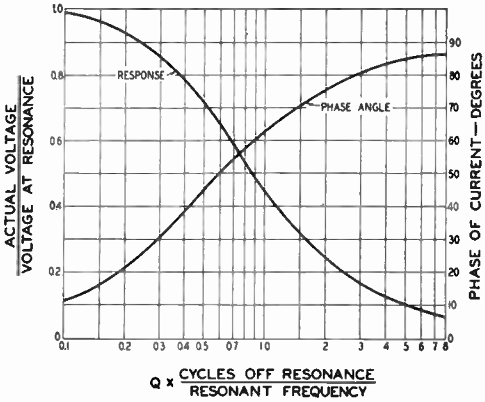
Mann, "An Inexpensive Sideband Filter," *QST*, March, 1949.

Rand, "The Little Slugger," *QST*, February, 1949.

Smith, "Premodulation Speech Clipping and Filtering," *QST*, February, 1946; "More on Speech Clipping," *QST*, March, 1947.

TUNED-CIRCUIT RESPONSE

The graph below gives the response and phase angle of a high-Q parallel-tuned circuit.



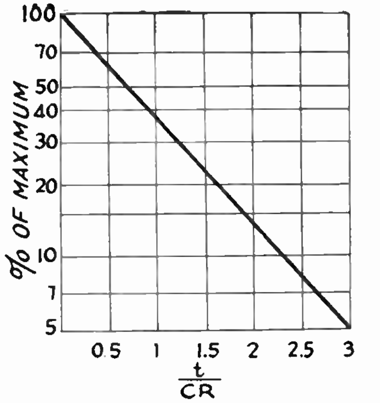
Circuit Q is equal to

$$2\pi fRC \quad \text{or} \quad \frac{R}{2\pi fL}$$

where L and C are the inductance and capacitance at the resonant frequency, f, and R is the parallel resistance across the circuit. The curves above become more accurate as the circuit Q is higher, but the error is not especially great for values as low as $Q = 10$

VOLTAGE DECAY IN *RC* CIRCUITS

The accompanying chart enables calculation of the instantaneous voltage across the terminals of a condenser discharging through a resistance. The voltage is given in terms of percentage of the voltage to which the condenser is initially charged. To obtain the voltage-decay time in seconds, multiply the factor (t/CR) by the time constant of the resistor-condenser circuit.



Example: A 0.01-μfd. condenser is charged to 150 volts and then allowed to discharge through a 0.1-megohm resistor. How long will it take the voltage to fall to 10 volts? In percentage, 10/150 = 6.7%. From the chart, the factor corresponding to 6.7% is 2.7. The time constant of the circuit is equal to $CR = 0.01 \times 0.1 = 0.001$. The time is therefore $2.7 \times 0.001 = 0.0027$ second, or 2.7 milliseconds.

Example: An *RC* circuit is desired in which the voltage will fall to 50% of the initial value in 0.1 second. From the chart, $t/CR = 0.7$ at the 50%-voltage point. Therefore $CR = t/0.7 = 0.1/0.7 = 1.43$. Any combination of resistance and capacitance whose product (*R* in megohms and *C* in microfarads) is equal to 1.43 can be used; for example, *C* could be 1 μfd. and *R* 1.43 megohms.

GERMANIUM CRYSTAL DIODES

Type	Max. Inverse Volts	Peak Rectif'd Ma.	Max. Surge Ma.	Max. Reverse μ-Amp.	Max. Average Ma.	Freq. Range Mc.
1N34	60	150	500	50 @ 10 V. 800 @ 50 V.	40	0–100
1N35[1]	50	60	100	10 @ 10 V.	22.5	0–100
1N38	100	150	500	6 @ 3 V. 625 @ 100 V.	40	0–100
1N39	200	150	500	200 @ 100 V. 800 @ 200 V.	40	0–100
1N40[2]	25	60	100	50 @ 10 V.	22.5	0–100
1N41[2]	25	60	100	50 @ 10 V.	22.5	0–100
1N42[2]	50	60	100	6 @ 3 V. 625 @ 100 V.	22.5	0–100
1N48 G5[3]	85	150	400	833 @ 50 V.	50	—
1N51 G5C[3]	50	100	300	1667 @ 50 V.	25	—
1N52 G5D[3]	85	150	400	150 @ 50 V.	50	—
1N54	35	150	500	10 @ 10 V.	40	0–100
1N55	150	150	500	300 @ 100 V. 800 @ 150 V.	40	0–100
1N56	40	200	1000	300 @ 30 V.	50	0–100
1N57	80	150	500	500 @ 75 V.	40	0–100
1N58	100	150	500	800 @ 100 V.	40	0–100
1N63 G5E[3]	125	150	400	50 @ 50 V.	50	—
1N64 G5F[3]	20	Specially designed for use as second detector in Tel. receivers.				—
1N65 G5G[3]	85	150	400	200 @ 50 V.	50	—
G7[3]	5	U.H.F. receiver mixer diode. Sensitivity 4–8 μv. Noise index 2.5–5 μv.				100–1000

Ratings given are for individual diodes. Average life is over 10,000 hours. Ambient temperature range for all types — $-50°$ C. to $+75°$ C. Average shunt capacitance — 0.8 $\mu\mu$fd.

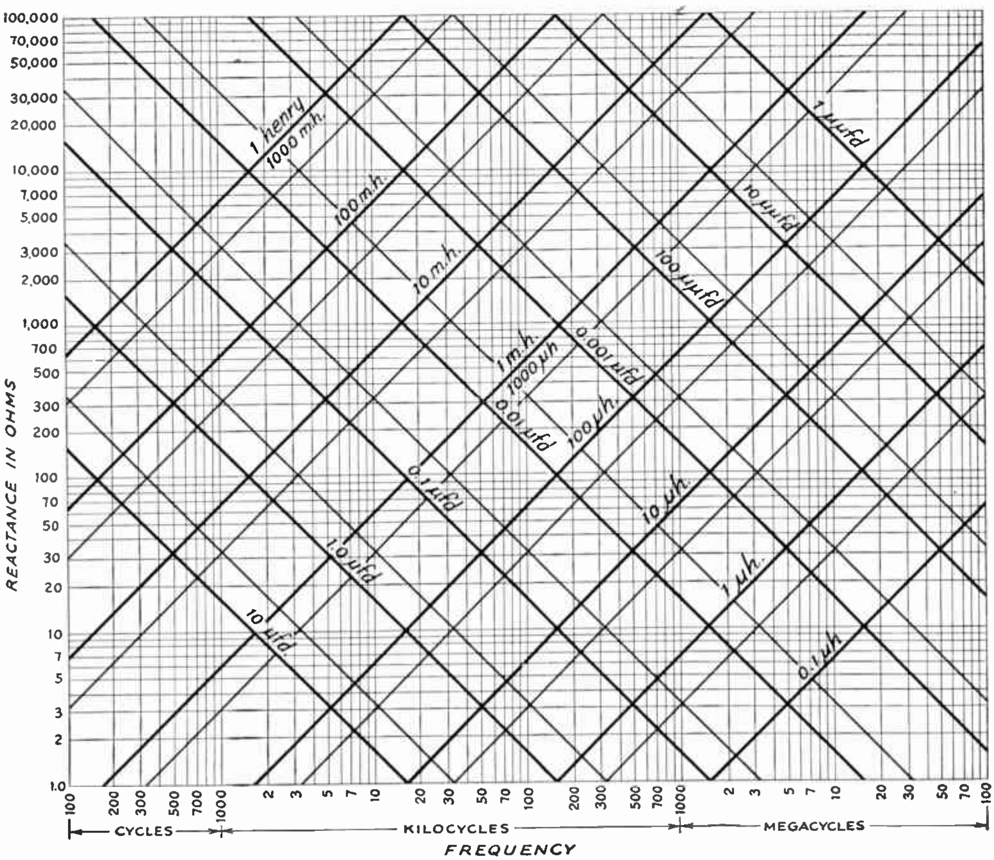
[1] Matched dual diode. [2] Unit has four matched diodes. [3] Manufactured by GE. Other units by Sylvania.

MINIATURE SELENIUM RECTIFIERS

Manufacturer	Type Number	Max. A.C. Volts	Peak Inverse Volts	Peak Current Ma.	Max. R.M.S. Ma.	Max. D.C. Output Ma.	Rectifier Service
Federal Telephone and Radio Corporation	402D3200	117	380	—	—	50	Half-Wave
"	402D2788# 402D3150A	117	380	900	220	75	Half-Wave
"	403D2625 403D2625A	117	380	1200	325	100	Half-Wave
"	402D3151	18	—	—	—	100	Half-Wave
"	402D3239A	160	—	—	—	75	Doubler
"	403D3240A	160	—	—	—	100	Doubler
General Electric Co.	6RS5GH2	117	380	650	163	65	Half-Wave
"	6RS5GH1	117	380	750	187	75	Half-Wave
Radio Receptor Company, Inc.	5L1	117	380	—	—	75	Half-Wave
"	5M1	117	380	—	—	100	Half-Wave

Circular plates—discontinued.

INDUCTIVE AND CAPACITIVE REACTANCE *VS.* FREQUENCY CHART



By use of the chart above, the approximate reactance of any capacitance from 1.0 μμfd. to 10 μfd. at any frequency from 100 cycles to 100 megacycles, or the reactance of any inductance from 0.1 μh. to 1.0 henry, can be read directly. Intermediate values can be estimated by interpolation In making interpolations, remember that the rate of change between lines is logarithmic. Use the frequency or reactance scales as a guide in estimating intermediate values on the capacitance or inductance scales.

This chart also can be used to find the approximate resonance frequencies of *LC* combinations, or the frequency to which a given coil-and-condenser combination will tune. First locate the respective slanting lines for the capacitance and inductance. The point where they intersect, i.e., where the reactances are equal, is the resonant frequency (projected downward and read on the frequency scale).

ELECTRICAL CONDUCTIVITY OF METALS

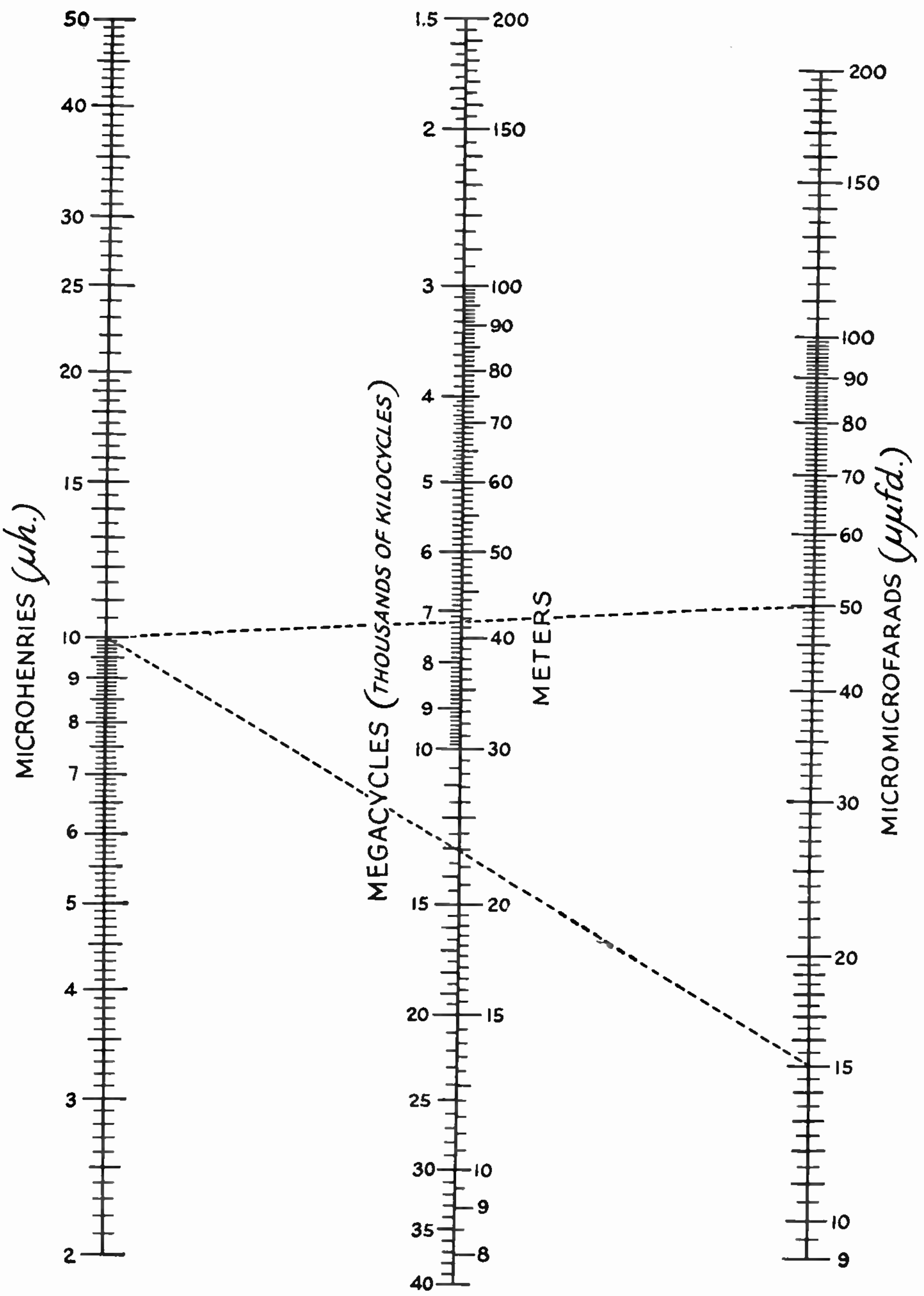
	Relative Conductivity [1]	*Temp. Coef.* [2] *of Resistance*
Aluminum (2S; pure)	59	0.0049
Aluminum (alloys):		
Soft-annealed	45–50	
Heat-treated	30–45	
Brass	28	0.002–0.007
Cadmium	19	
Chromium	55	
Climax	1.83	
Cobalt	16.3	
Constantin	3.24	0.00002
Copper (hard drawn)	89.5	0.004
Copper (annealed)	100	
Everdur	6	
German Silver (18%)	5.3	0.00019
Gold	65	
Iron (pure)	17.7	0.006
Iron (cast)	2–12	
Iron (wrought)	11.4	
Lead	7	0.0041
Manganin	3.7	0.00002
Mercury	1.66	0.00089
Molybdenum	33.2	0.0033
Monel	4	0.0019
Nichrome	1.45	0.00017
Nickel	12–16	0.005
Phosphor Bronze	36	0.004
Platinum	15	
Silver	106	0.004
Steel	3–15	
Tin	13	0.0042
Tungsten	28.9	0.0045
Zinc	28.2	0.0035

[1] At 20° C., based on copper as 100. [2] Per °C. at 20° C.

Approximate relations:

An increase of 1 in A. W. G. or B. & S. wire size increases resistance 25%.
An increase of 2 increases resistance 60%.
An increase of 3 increases resistance 100%.
An increase of 10 increases resistance 10 times.

INDUCTANCE, CAPACITANCE AND FREQUENCY CHART — 1.5–40 MC.



This chart may be used to find the values of inductance and capacitance required to resonate at any given frequency in the medium- or high-frequency ranges; or, conversely, to find the frequency to which any given coil-condenser combination will tune. In the example shown by the dashed lines, a condenser has a minimum capacitance of 15 μμfd. and a maximum capacitance of 50 μμfd. If it is to be used with a coil of 10-μh. inductance, what frequency range will be covered? The straightedge is connected between 10 on the left-hand scale and 15 on the right, giving 13 Mc. as the high-frequency limit. Keeping the straightedge at 10 on the left-hand scale, the other end is swung to 50 on the right-hand scale, giving a low-frequency limit of 7.1 Mc. The tuning range would, therefore, be from 7.1 Mc. to 13 Mc., or 7100 kc. to 13,000 kc. The center scale also serves to convert frequency to wavelength.

The range of the chart can be extended by multiplying each of the scales by 0.1 or 10. In the example above, if the capacitances are 150 and 500 μμfd. and the inductance 100 μh., the range becomes approximately 231 to 122 meters or 0.7 to 1.3 Mc. Alternatively, 1.5 to 5 μμfd. and 1 μh. will give a range of approximately 71 to 130 Mc.

COPPER-WIRE TABLE

Gauge No. B. & S.	Diam. in Mils[1]	Circular Mil Area	Turns per Linear Inch[2]				Turns per Square Inch[2]			Feet per Lb.		Ohms per 1000 ft. 25° C.	Current Carrying Capacity at 1500 C.M. per Amp.[3]	Diam. in mm.	Nearest British S.W.G. No.
			Enamel	S.S.C.	D.S.C. or S.C.C.	D.C.C.	S.C.C.	Enamel S.C.C.	D.C.C.	Bare	D.C.C.				
1	289.3	83690	—	—	—	—	—	—	—	3.947	—	.1264	55.7	7.348	1
2	257.6	66370	—	—	—	—	—	—	—	4.977	—	.1593	44.1	6.544	3
3	229.4	52640	—	—	—	—	—	—	—	6.276	—	.2009	35.0	5.827	4
4	204.3	41740	—	—	—	—	—	—	—	7.914	—	.2533	27.7	5.189	5
5	181.9	33100	—	—	—	—	—	—	—	9.980	—	.3195	22.0	4.621	7
6	162.0	26250	—	—	—	—	—	—	—	12.58	—	.4028	17.5	4.115	8
7	144.3	20820	—	—	—	—	—	—	—	15.87	—	.5080	13.8	3.665	9
8	128.5	16510	7.6	—	7.4	7.1	—	—	—	20.01	19.6	.6405	11.0	3.264	10
9	114.4	13090	8.6	—	8.2	7.8	—	—	—	25.23	24.6	.8077	8.7	2.906	11
10	101.9	10380	9.6	—	9.3	8.9	87.5	84.8	80.0	31.82	30.9	1.018	6.9	2.588	12
11	90.74	8234	10.7	—	10.3	9.8	110	105	97.5	40.12	38.8	1.284	5.5	2.305	13
12	80.81	6530	12.0	—	11.5	10.9	136	131	121	50.59	48.9	1.619	4.4	2.053	14
13	71.96	5178	13.5	—	12.8	12.0	170	162	150	63.80	61.5	2.042	3.5	1.828	15
14	64.08	4107	15.0	—	14.2	13.8	211	198	183	80.44	77.3	2.575	2.7	1.628	16
15	57.07	3257	16.8	—	15.8	14.7	262	250	223	101.4	97.3	3.247	2.2	1.450	17
16	50.82	2583	18.9	18.9	17.9	16.4	321	306	271	127.9	119	4.094	1.7	1.291	18
17	45.26	2048	21.2	21.2	19.9	18.1	397	372	329	161.3	150	5.163	1.3	1.150	18
18	40.30	1624	23.6	23.6	22.0	19.8	493	454	399	203.4	188	6.510	1.1	1.024	19
19	35.89	1288	26.4	26.4	24.4	21.8	592	553	479	256.5	237	8.210	.86	.9116	20
20	31.96	1022	29.4	29.4	27.0	23.8	775	725	625	323.4	298	10.35	.68	.8118	21
21	28.46	810.1	33.1	32.7	29.8	26.0	940	895	754	407.8	370	13.05	.54	.7230	22
22	25.35	642.4	37.0	36.5	34.1	30.0	1150	1070	910	514.2	461	16.46	.43	.6438	23
23	22.57	509.5	41.3	40.6	37.6	31.6	1400	1300	1080	648.4	584	20.76	.34	.5733	24
24	20.10	404.0	46.3	45.3	41.5	35.6	1700	1570	1260	817.7	745	26.17	.27	.5106	25
25	17.90	320.4	51.7	50.4	45.6	38.6	2060	1910	1510	1031	903	33.00	.21	.4547	26
26	15.94	254.1	58.0	55.6	50.2	41.8	2500	2300	1750	1300	1118	41.62	.17	.4049	27
27	14.20	201.5	64.9	61.5	55.0	45.0	3030	2780	2020	1639	1422	52.48	.13	.3606	29
28	12.64	159.8	72.7	68.6	60.2	48.5	3670	3350	2310	2067	1759	66.17	.11	.3211	30
29	11.26	126.7	81.6	74.8	65.4	51.8	4300	3900	2700	2607	2207	83.44	.084	.2859	31
30	10.03	100.5	90.5	83.3	71.5	55.5	5040	4660	3020	3287	2534	105.2	.067	.2546	33
31	8.928	79.70	101	92.0	77.5	59.2	5920	5280	—	4145	2768	132.7	.053	.2268	34
32	7.950	63.21	113	101	83.6	62.6	7060	6250	—	5227	3137	167.3	.042	.2019	36
33	7.080	50.13	127	110	90.3	66.3	8120	7360	—	6591	4697	211.0	.033	.1798	37
34	6.305	39.75	143	120	97.0	70.0	9600	8310	—	8310	6168	266.0	.026	.1601	38
35	5.615	31.52	158	132	104	73.5	10900	8700	—	10480	6737	335.0	.021	.1426	38–39
36	5.000	25.00	175	143	111	77.0	12200	10700	—	13210	7877	423.0	.017	.1270	39–40
37	4.453	19.83	198	154	118	80.3	—	—	—	16660	9309	533.4	.013	.1131	41
38	3.965	15.72	224	166	126	83.6	—	—	—	21010	10666	672.6	.010	.1007	42
39	3.531	12.47	248	181	133	86.6	—	—	—	26500	11907	848.1	.008	.0897	43
40	3.145	9.88	282	194	140	89.7	—	—	—	33410	14222	1069	.006	.0799	44

[1] A mil is 1/1000 (one-thousandth) of an inch.
[2] The figures given are approximate only, since the thickness of the insulation varies with different manufacturers.
[3] The current-carrying capacity at 1000 C.M. per ampere is equal to the circular-mil area (Column 3) divided by 1000.

STANDARD METAL GAUGES

Gauge No.	American or B. & S.[1]	U. S. Standard[2]	Birmingham or Stubs[3]
1	.2893	.28125	.300
2	.2576	.265625	.284
3	.2294	.25	.259
4	.2043	.234375	.238
5	.1819	.21875	.220
6	.1620	.203125	.203
7	.1443	.1875	.180
8	.1285	.171875	.165
9	.1144	.15625	.148
10	.1019	.140625	.134
11	.09074	.125	.120
12	.08081	.109375	.109
13	.07196	.09375	.095
14	.06408	.078125	.083
15	.05707	.0703125	.072
16	.05082	.0625	.065
17	.04526	.05625	.058
18	.04030	.05	.049
19	.03589	.04375	.042
20	.03196	.0375	.035
21	.02846	.034375	.032
22	.02535	.03125	.028
23	.02257	.028125	.025
24	.02010	.025	.022
25	.01790	.021875	.020
26	.01594	.01875	.018
27	.01420	.0171875	.016
28	.01264	.015625	.014
29	.01126	.0140625	.013
30	.01003	.0125	.012
31	.008928	.0109375	.010
32	.007950	.01015625	.009
33	.007080	.009375	.008
34	.006350	.00859375	.007
35	.005615	.0078125	.005
36	.005000	.00703125	.004
37	.004453	.006640626	....
38	.003965	.00625	....
39	.003531	......	....
40	.003145	......	....

[1] Used for aluminum, copper, brass and nonferrous alloy sheets, wire and rods.
[2] Used for iron, steel, nickel and ferrous alloy sheets, wire and rods.
[3] Used for seamless tubes; also by some manufacturers for copper and brass.

MUSICAL SCALE

Approximate frequencies of notes of the musical scale, based on A-440.

	Note	Frequency
(Bottom Octave)	A-1	28
	A#-1	29
	B-1	31
Third octave below Middle C	Co	33
	C#o	35
	Do	37
	D#o	39
	Eo	41
	Fo	44
	F#o	46
	Go	49
	G#o	52
	Ao	55
	A#o	58
	Bo	62
Second octave below Middle C	C1	65
	C#1	69
	D1	73
	D#1	78
	E1	82
	F1	87
	F#1	93
	G1	98
	G#1	104
	A1	110
	A#1	117
	B1	123
First octave below Middle C	C2	131
	C#2	139
	D2	147
	D#2	156
	E2	165
	F2	175
	F#2	185
	G2	196
	G#2	208
	A2	220
	A#2	233
	B2	247
Middle C — First octave above Middle C	C3	262
	C#3	277
	D3	294
	D#3	311
	E3	330
	F3	349
	F#3	370
	G3	392
	G#3	415
	A3	440
	A#3	466
	B3	494
Second octave above Middle C	C4	523
	C#4	554
	D4	587
	D#4	622
	E4	659
	F4	698
	F#4	740
	G4	784
	G#4	831
	A4	880
	A#4	932
	B4	988
Third octave above Middle C	C5	1047
	C#5	1109
	D5	1175
	D#5	1245
	E5	1319
	F5	1397
	F#5	1480
	G5	1568
	G#5	1661
	A5	1760
	A#5	1865
	B5	1976
Fourth octave above Middle C	C6	2093
	C#6	2217
	D6	2349
	D#6	2489
	E6	2637
	F6	2794
	F#6	2960
	G6	3136
	G#6	3322
	A6	3520
	A#6	3729
	B6	3951
	C7	4186

LETTER SYMBOLS FOR VACUUM-TUBE NOTATION

Grid potential	E_g, e_g	Mutual conductance	g_m
Grid current	I_g, i_g	Amplification factor	μ
Grid conductance	g_g	Filament terminal voltage	E_f
Grid resistance	r_g	Filament current	I_f
Grid bias voltage	E_c	Grid-plate capacitance	C_{gp}
Plate potential	E_p, e_p	Grid-cathode capacitance	C_{gk}
Plate current	I_b, I_p, i_p	Plate-cathode capacitance	C_{pk}
Plate conductance	g_p	Grid capacitance (input)	C_g
Plate resistance	r_p	Plate capacitance (output)	C_p
Plate supply voltage	E_b		
Cathode current	I_c		
Emission current	I_s		

NOTE. — Small letters refer to instantaneous values.

GREEK ALPHABET

Greek Letter	*Greek Name*	*English Equivalent*
Α α	Alpha	a
Β β	Beta	b
Γ γ	Gamma	g
Δ δ	Delta	d
Ε ε	Epsilon	e
Ζ ζ	Zeta	z
Η η	Eta	é
Θ θ	Theta	th
Ι ι	Iota	i
Κ κ	Kappa	k
Λ λ	Lambda	l
Μ μ	Mu	m
Ν ν	Nu	n
Ξ ξ	Xi	x
Ο ο	Omicron	ŏ
Π π	Pi	p
Ρ ρ	Rho	r
Σ σ	Sigma	s
Τ τ	Tau	t
Υ υ	Upsilon	u
Φ φ	Phi	ph
Χ χ	Chi	ch
Ψ ψ	Psi	ps
Ω ω	Omega	ō

THE R-S-T SYSTEM

READABILITY

1 — Unreadable.
2 — Barely readable, occasional words distinguishable.
3 — Readable with considerable difficulty.
4 — Readable with practically no difficulty.
5 — Perfectly readable.

SIGNAL STRENGTH

1 — Faint signals, barely perceptible.
2 — Very weak signals.
3 — Weak signals.
4 — Fair signals.
5 — Fairly good signals.
6 — Good signals.
7 — Moderately strong signals.
8 — Strong signals.
9 — Extremely strong signals.

TONE

1 — Extremely rough hissing note.
2 — Very rough a.c. note, no trace of musicality.
3 — Rough low-pitched a.c. note, slightly musical.
4 — Rather rough a.c. note, moderately musical.
5 — Musically-modulated note.
6 — Modulated note, slight trace of whistle.
7 — Near d.c. note, smooth ripple.
8 — Good d.c. note, just a trace of ripple.
9 — Purest d.c. note.

If the signal has the characteristic steadiness of crystal control, add the letter X to the RST report. If there is a chirp, the letter C may be added to so indicate. Similarly for a click, add K. The above reporting system is used on both c.w. and voice, leaving out the "tone" report on voice.

Q SIGNALS

Given below are a number of Q signals whose meanings most often need to be expressed with brevity and clearness in amateur work. (Q abbreviations take the form of questions only when each is sent followed by a question mark.)

QRG Will you tell me my exact frequency (or that of.......)? Your exact frequency (or that of......) is......kc.
QRH Does my frequency vary? Your frequency varies.
QRI How is the tone of my transmission? The tone of your transmission is..... (1. Good; 2. Variable; 3. Bad).
QRK What is the readability of my signals (or those of......)? The readability of your signals (or those of.....) is..... (1. Unreadable; 2. Readable now and then; 3. Readable but with difficulty; 4. Readable; 5. Perfectly readable).
QRL Are you busy? I am busy (or I am busy with ......). Please do not interfere.
QRM Are you being interfered with? I am interfered with.
QRN Are you troubled by static? I am being troubled by static.
QRQ Shall I send faster? Send faster (......words per min.).
QRS Shall I send more slowly? Send more slowly (.... w.p.m.).
QRT Shall I stop sending? Stop sending.
QRU Have you anything for me? I have nothing for you.
QRV Are you ready? I am ready.
QRW Shall I tell.....that you are calling him on .....kc.? Please inform.....that I am calling him on.....kc.
QRX When will you call me again? I will call you again at......hours (on........kc.).
QRZ Who is calling me? You are being called by..... (on......kc.).
QSA What is the strength of my signals (or those of ......)? The strength of your signals (or those of.....) is....... (1. Scarcely perceptible; 2. Weak; 3. Fairly good; 4. Good; 5. Very good).
QSB Are my signals fading? Your signals are fading.
QSD Is my keying defective? Your keying is defective.
QSG Shall I send.....messages at a time? Send..... messages at a time.
QSL Can you acknowledge receipt? I am acknowledging receipt.
QSM Shall I repeat the last message which I sent you, or some previous message? Repeat the last message which you sent me [or message(s) number(s).....].
QSO Can you communicate with....direct or by relay? I can communicate with.....direct (or by relay through.....).
QSP Will you relay to.....? I will relay to.....
QSV Shall I send a series of Vs on this frequency (or ....kc.)? Send a series of Vs on this frequency (or.....kc.).
QSW Will you send on this frequency (or on....kc.)? I am going to send on this frequency (or on .....kc.).
QSX Will you listen to.....on.....kc.? I am listening to......on.....kc.
QSY Shall I change to transmission on another frequency? Change to transmission on another frequency (or on....kc.).
QSZ Shall I send each word or group more than once? Send each word or group twice (or....times).
QTA Shall I cancel message number....as if it had not been sent? Cancel message number.....as if it had not been sent.
QTB Do you agree with my counting of words? I do not agree with your counting of words; I will repeat the first letter or digit of each word or group.
QTC How many messages have you to send? I have.... messages for you (or for.....).
QTH What is your location? My location is.....
QTR What is the exact time? The time is......

Special abbreviations adopted by ARRL:
QST General call preceding a message addressed to all amateurs and ARRL members. This is in effect "CQ ARRL."
QRRR Official ARRL "land SOS." A distress call for emergency use only by a station in an emergency situation.

ABBREVIATIONS FOR C.W. WORK

Abbreviations help to cut down unnecessary transmission. However, make it a rule not to abbreviate unnecessarily when working an operator of unknown experience.

AA	All after
AB	All before
ABT	About
ADR	Address
AGN	Again
ANT	Antenna
BCI	Broadcast interference
BCL	Broadcast listener
BK	Break; break me; break in
BN	All between; been
B4	Before
C	Yes
CFM	Confirm; I confirm
CK	Check
CL	I am closing my station; call
CLD-CLG	Called; calling
CUD	Could
CUL	See you later
CUM	Come
CW	Continuous wave
DLD-DLVD	Delivered
DX	Distance
ECO	Electron-coupled oscillator
FB	Fine business; excellent
GA	Go ahead (or resume sending)
GB	Good-by
GBA	Give better address
GE	Good evening
GG	Going
GM	Good morning
GN	Good night
GND	Ground
GUD	Good
HI	The telegraphic laugh; high
HR	Here; hear
HV	Have
HW	How
LID	A poor operator
MILS	Milliamperes
MSG	Message; prefix to radiogram
N	No
ND	Nothing doing
NIL	Nothing; I have nothing for you
NR	Number
NW	Now; I resume transmission
OB	Old boy
OM	Old man
OP-OPR	Operator
OSC	Oscillator
OT	Old timer; old top
PBL	Preamble
PSE-PLS	Please
PWR	Power
PX	Press
R	Received solid; all right; OK; are
RAC	Rectified alternating current
RCD	Received
REF	Refer to; referring to; reference
RPT	Repeat; I repeat
SED	Said
SEZ	Says
SIG	Signature; signal
SINE	Operator's personal initials or nickname
SKED	Schedule
SRI	Sorry
SVC	Service; prefix to service message
TFC	Traffic
TMW	Tomorrow
TNX-TKS	Thanks
TT	That
TU	Thank you
TXT	Text
UR-URS	Your; you're; yours
VFO	Variable-frequency oscillator
VY	Very
WA	Word after
WB	Word before
WD-WDS	Word; words
WKD-WKG	Worked; working
WL	Well; will
WUD	Would
WX	Weather
XMTR	Transmitter
XTAL	Crystal
YF (XYL)	Wife
YL	Young lady
73	Best regards
88	Love and kisses

W PREFIXES BY STATES

Alabama	W4
Arizona	W7
Arkansas	W5
California	W6
Colorado	WØ
Connecticut	W1
Delaware	W3
District of Columbia	W3
Florida	W4
Georgia	W4
Idaho	W7
Illinois	W9
Indiana	W9
Iowa	WØ
Kansas	WØ
Kentucky	W4
Louisiana	W5
Maine	W1
Maryland	W3
Massachusetts	W1
Michigan	W8
Minnesota	WØ
Mississippi	W5
Missouri	WØ
Montana	W7
Nebraska	WØ
Nevada	W7
New Hampshire	W1
New Jersey	W2
New Mexico	W5
New York	W2
North Carolina	W4
North Dakota	WØ
Ohio	W8
Oklahoma	W5
Oregon	W7
Pennsylvania	W3
Rhode Island	W1
South Carolina	W4
South Dakota	WØ
Tennessee	W4
Texas	W5
Utah	W7
Vermont	W1
Virginia	W4
Washington	W7
West Virginia	W8
Wisconsin	W9
Wyoming	W7

INTERNATIONAL PREFIXES

Below is the list of prefixes assigned to the countries of the world by the 1947 International Telecommunications Conference at Atlantic City. These assignments became effective on January 1, 1949.

AAA-ALZ	United States of America
AMA-AOZ	(Not allocated)
APA-ASZ	Pakistan
ATA-AWZ	India
AXA-AXZ	Commonwealth of Australia
AYA-AZZ	Argentina Republic
BAA-BZZ	China
CAA-CEZ	Chile
CFA-CKZ	Canada
CLA-CMZ	Cuba
CNA-CNZ	Morocco
COA-COZ	Cuba
CPA-CPZ	Bolivia
CQA-CRZ	Portuguese Colonies
CSA-CUZ	Portugal
CVA-CXZ	Uruguay
CYA-CZZ	Canada
DAA-DMZ	Germany
DNA-DQZ	Belgian Congo
DRA-DTZ	Bielorussian Soviet Socialist Republic
DUA-DZZ	Republic of the Philippines
EAA-EHZ	Spain
EIA-EJZ	Ireland
EKA-EKZ	Union of Soviet Socialist Republics
ELA-ELZ	Republic of Liberia
EMA-EOZ	Union of Soviet Socialist Republics
EPA-EQZ	Iran
ERA-ERZ	Union of Soviet Socialist Republics
ESA-ESZ	Estonia
ETA-ETZ	Ethiopa
EUA-EZZ	Union of Soviet Socialist Republics
FAA-FZZ	France and Colonies and Protectorates
GAA-GZZ	Great Britain
HAA-HAZ	Hungary
HBA-HBZ	Switzerland
HCA-HDZ	Ecuador
HEA-HEZ	Switzerland
HFA-HFZ	Poland
HGA-HGZ	Hungary
HHA-HHZ	Republic of Haiti
HIA-HIZ	Dominican Republic
HJA-HKZ	Republic of Colombia
HLA-HMZ	Korea
HNA-HNZ	Iraq
HOA-HPZ	Republic of Panama
HQA-HRZ	Republic of Honduras
HSA-HSZ	Siam
HTA-HTZ	Nicaragua
HUA-HUZ	Republic of El Salvador
HVA-HVZ	Vatican City State
HWA-HYZ	France and Colonies and Protectorates
HZA-HZZ	Kingdom of Saudi Arabia
IAA-IZZ	Italy and Colonies
JAA-JSZ	Japan
JTA-JVZ	Mongolian People's Republic
JWA-JXZ	Norway
JYA-JZZ	(Not allocated)
KAA-KZZ	United States of America
LAA-LNZ	Norway
LOA-LWZ	Argentina Republic
LXA-LXZ	Luxembourg
LYA-LYZ	Lithuania
LZA-LZZ	Bulgaria
MAA-MZZ	Great Britain
NAA-NZZ	United States of America
OAA-OCZ	Peru
ODA-ODZ	Republic of Lebanon
OEA-OEZ	Austria
OFA-OJZ	Finland
OKA-OMZ	Czechoslovakia
ONA-OTZ	Belgium and Colonies
OUA-OZZ	Denmark
PAA-PIZ	Netherlands
PJA-PJZ	Curacao
PKA-POZ	Netherlands Indies
PPA-PYZ	Brazil
PZA-PZZ	Surinam
QAA-QZZ	(Service abbreviations)
RAA-RZZ	Union of Soviet Socialist Republics
SAA-SMZ	Sweden
SNA-SRZ	Poland
SSA-SUZ	Egypt
SVA-SZZ	Greece
TAA-TCZ	Turkey
TDA-TDZ	Guatemala
TEA-TEZ	Costa Rica
TFA-TFZ	Iceland
TGA-TGZ	Guatemala
THA-THZ	France and Colonies and Protectorates
TIA-TIZ	Costa Rica
TJA-TZZ	France and Colonies and Protectorates
UAA-UQZ	Union of Soviet Socialist Republics
URA-UTZ	Ukrainian Soviet Socialist Republic
UUA-UZZ	Union of Soviet Socialist Republics
VAA-VGZ	Canada
VHA-VNZ	Commonwealth of Australia
VOA-VOZ	Newfoundland
VPA-VSZ	British Colonies and Protectorates
VTA-VWZ	India
VXA-VYZ	Canada
VZA-VZZ	Commonwealth of Australia
WAA-WZZ	United States of America
XAA-XIZ	Mexico
XJA-XOZ	Canada
XPA-XPZ	Denmark
XQA-XRZ	Chile
XSA-XSZ	China
XTA-XWZ	France and Colonies and Protectorates
XXA-XXZ	Portuguese Colonies
XYA-XZZ	Burma
YAA-YAZ	Afghanistan
YBA-YHZ	Netherlands Indies
YIA-YIZ	Iraq
YJA-YJZ	New Hebrides
YKA-YKZ	Syria
YLA-YLZ	Latvia
YMA-YMZ	Turkey
YNA-YNZ	Nicaragua
YOA-YRZ	Roumania
YSA-YSZ	Republic of El Salvador
YTA-YUZ	Yugoslavia
YVA-YYZ	Venezuela
YZA-YZZ	Yugoslavia
ZAA-ZAZ	Albania
ZBA-ZJZ	British Colonies and Protectorates
ZKA-ZMZ	New Zealand
ZNA-ZOZ	British Colonies and Protectorates
ZPA-ZPZ	Paraguay
ZQA-ZQZ	British Colonies and Protectorates
ZRA-ZUZ	Union of South Africa
ZVA-ZZZ	Brazil
2AA-2ZZ	Great Britain
3AA-3AZ	Principality of Monaco
3BA-3FZ	Canada
3GA-3GZ	Chile
3HA-3UZ	China
3VA-3VZ	France and Colonies and Protectorates
3WA-3XZ	(Not allocated)
3YA-3YZ	Norway
3ZA-3ZZ	Poland
4AA-4CZ	Mexico
4DA-4IZ	Republic of the Philippines
4JA-4LZ	Union of Soviet Socialist Republics
4MA-4MZ	Venezuela
4NA-4OZ	Yugoslavia
4PA-4SZ	British Colonies and Protectorates
4TA-4TZ	Peru
4UA-4UZ	United Nations
4VA-4VZ	Republic of Haiti
4WA-4WZ	Yemen
4XA-4ZZ	(Not allocated)
5AA-5ZZ	(Not allocated)
6AA-6ZZ	(Not allocated)
7AA-7ZZ	(Not allocated)
8AA-8ZZ	(Not allocated)
9AA-9ZZ	(Not allocated)

A.R.R.L. COUNTRIES LIST

Official List for ARRL DX Contest and the Postwar DXCC

Country	Prefix
Aden and Socotra Island	VS9
Afghanistan	YA
Alaska	KL7
Albania	ZA
Aldabra Islands	
Algeria	FA
Andaman Ids. and Nicobar Ids	VU
Andorra	PX
Anglo-Egyptian Sudan	ST
Angola	CR6
Antarctica	
Argentina	LU
Ascension Island	ZD8
Australia (including Tasmania)	VK
Austria	(MB9) . . OE
Azores Islands	CT2
Bahama Islands	VP7
Bahrein Island	VU7
Baker Island, Howland Island and Am. Phoenix Islands	KB6
Balearic Islands	EA6
Barbados	VP6
Basutoland	ZS8
Bechuanaland	ZS9
Belgian Congo	OQ
Belgium	ON
Bermuda Islands	VP9
Bhutan	
Bolivia	CP
Bonin Islands and Volcano Islands (e.g., Iwo Jima)	
Borneo, British North	VS4
Borneo, Netherlands	PK5
Brazil	PY
Brunei	VS5
Bulgaria	LZ
Burma	XZ
Cameroons, French	FE8
Canada	VE
Canal Zone	KZ5
Canary Islands	EA8
Cape Verde Islands	CR4
Caroline Islands	KC6
Cayman Islands	VP5
Celebes and Molucca Islands	PK6
Ceylon	VS7
Chagos Islands	VQ8
Channel Islands	GC
Chile	CE
China	B, C
Christmas Island	ZC3
Clipperton Island	
Cocos Island	TI
Cocos Islands	ZC2
Colombia	HK
Comoro Islands	
Cook Islands	ZK1
Corsica	FC
Costa Rica	TI
Crete	SV
Cuba	CM-CO
Cyprus	(MD7) . . ZC4
Czechoslovakia	OK
Denmark	OZ
Dodecanese Islands (e.g., Rhodes)	SV5
Dominican Republic	HI
Easter Island	
Ecuador	HC
Egypt	(MD5) . . SU
Eire (Irish Free State)	EI
England	G
Eritrea	I6
Ethiopia	ET
Faeroes, The	OY
Falkland Islands	VP8
Fanning Island (Christmas Island)	VR3
Fiji Islands	VR2
Finland	OH
Formosa (Taiwan)	C3
France	F
French Equatorial Africa	FQ8
French India	FN
French Indo-China	FI8
French Oceania (e.g., Tahiti)	FO8
French West Africa	FF8
Fridtjof Nansen Land (Franz Josef Land)	UA1
Galapagos Islands	
Gambia	ZD3
Germany	D
Gibraltar	ZB2
Gilbert & Ellice Islands and Ocean Island	VR1
Goa (Portuguese India)	CR8
Gold Coast (and British Togoland)	ZD4
Greece	SV
Greenland	OX
Guadeloupe	FG8
Guantanamo Bay	KG4
Guatemala	TG
Guiana, British	VP3
Guiana, Netherlands (Surinam)	PZ
Guiana, French, and Inini	FY8
Guinea, Portuguese	CR5
Guinea, Spanish	
Haiti	HH
Hawaiian Islands	KH6
Heard Island	VK1
Honduras	HR
Honduras, British	VP1
Hong Kong	VS6
Hungary	HA
Iceland	TF
Ifni	
India	VU
Iran	EP-EQ
Iraq	YI
Ireland, Northern	GI
Isle of Man	GD
Israel	4X4
Italy	I
Jamaica	VP5
Jan Mayen Island	
Japan	J
Jarvis Island, Palmyra group (Christmas Island)	KP6
Java	PK
Johnston Island	KJ6
Kenya	VQ4
Kerguelen Islands	
Korea	HL
Kuwait	
Laccadive Islands	VU4
Lebanon	AR8
Leeward Islands	VP2
Liberia	EL
Libya	(MD1-2) . . LI
Liechtenstein	HE1
Luxembourg	LX
Macau	CR9
MacQuarie Island	VK1
Madagascar	FB8
Madeira Islands	CT3
Malaya	VS1, VS2
Maldive Islands	
Malta	ZB1
Manchuria	C9
Marianas Islands (Guam)	KG6
Marion Island (Prince Edward Island)	ZS
Marshall Islands	KX6
Martinique	FM8
Mauritius	VQ8
Mexico	XE
Midway Island	KM6
Miquelon and St. Pierre Islands	FP8
Monaco	CZ
Mongolian Republic (Outer)	
Morocco, French	CN
Morocco, Spanish	EA9
Mozambique	CR7
Nepal	
Netherlands	PA
Netherlands West Indies	PJ
New Caledonia	FK8
New Guinea, Netherlands	PK6
New Guinea, Territory of	VK9
New Hebrides	FU8, YJ
New Zealand	ZL
Nicaragua	YN
Nigeria	ZD2
Niue	ZK2
Norfolk Island	VK9
Norway	LA
Nyasaland	ZD6
Oman	(MP4) . . VS9
Pakistan	AP
Palau (Pelew) Islands	
Palestine	ZC6
Panama	HP
Papua Territory	VK9
Paraguay	ZP
Peru	OA
Philippine Islands	DU
Phoenix Islands (British)	VR1
Pitcairn Island	VR6
Poland	SP
Portugal	CT
Principe and Sao Thome Islands	
Puerto Rico	KP4
Reunion Island	FR8
Rhodesia, Northern	VQ2
Rhodesia, Southern	ZE
Rio de Oro	
Roumania	YR
Ryukyu Islands (e.g., Okinawa)	KR6
St. Helena	ZD7
Salvador	YS
Samoa, American	KS6
Samoa, Western	ZM
San Marino	M1
Sarawak	VS5
Sardinia	IS
Saudi Arabia (Hedjaz and Nejd)	HZ
Scotland	GM
Seychelles	VQ9
Siam	HS
Sierra Leone	ZD1
Sikkim	AC3
Solomon Islands	VR4
Somaliland, British	VQ6
Somaliland, French	FL8
Somaliland, Italian	(MD4)
South Georgia	VP8
South Orkney Islands	VP8
South Sandwich Islands	VP8
South Shetland Islands	VP8
Southwest Africa	ZS3
Soviet Union:	
European Russian Socialist Federated Soviet Republic	UA1-3-4-6
Asiatic Russian S.F.S.R.	UA9-0
Ukraine	UB5
White Russian Soviet Socialist Republic	UC
Azerbaijan	UD6
Georgia	UF6
Armenia	UG6
Turkoman	UH8
Uzbek	UI8
Tadzhik	UJ8
Kazakh	UL7
Kirghiz	UM8
Karelo-Finnish Republic	UN1
Moldavia	UO5
Lithuania	UP
Latvia	UQ
Estonia	UR
Spain	EA
Sumatra	PK4
Svalbard (Spitzbergen)	LA
Swan Island	KS4
Swaziland	ZS7
Sweden	SM
Switzerland	HB
Syria	YK
Tanganyika Territory	VQ3
Tangier Zone	EK
Tannu Tuva	
Tibet	AC4
Timor, Portuguese	CR10
Togoland, French	FD8
Tokelau (Union) Islands	
Tonga (Friendly) Islands	VR5
Trans-Jordan	ZC1
Trieste	
Trinidad and Tobago	VP4
Tristan da Cunha and Gough Island	ZD9
Tunisia	3V8
Turkey	TA
Turks and Caicos Islands	VP5
Uganda	VQ5
Union of South Africa	ZS
United States of America	W, K
Uruguay	CX
Vatican City	HV
Venezuela	YV
Virgin Islands	KV4
Wake Island	KW6
Wales	GW
Windward Islands	VP2
Wrangel Islands	
Yemen	
Yugoslavia	YT-YU
Zanzibar	VQ1

NOTE: Prefixes in parentheses are used by occupation forces.

INTERNATIONAL AMATEUR PREFIXES

To make possible identification of calls heard on the air, the international telecommunications conferences assign to each nation certain alphabetical blocks, from which *all* classes of stations are assigned prefixes. The following prefixes are used by amateurs:

AC3	Sikkim
AC4	Tibet
AP	Pakistan
AR8	Lebanon
C	China (unofficial)
C3	Formosa
C9	Manchuria
CE	Chile
CM, CO	Cuba
CN	Morocco, French
CP	Bolivia
CR4	Cape Verde Islands
CR5	Guinea, Portuguese
CR6	Angola
CR7	Mozambique
CR8	Goa (Portuguese India)
CR9	Macau
CR10	Timor, Portuguese
CT1	Portugal
CT2	Azores Islands
CT3	Madeira Islands
CX	Uruguay
CZ	Monaco
DL	Germany
DU	Philippine Islands
EA	Spain
EA6	Balearic Islands
EA8	Canary Islands
EA9	Morocco, Spanish
EI	Eire (Irish Free State)
EK	Tangier Zone
EL	Liberia
EP, EQ	Iran (Persia)
ET	Ethiopia
F	France
FA	Algeria
FB8	Madagascar
FC	Corsica
FD8	Togoland, French
FE8	Cameroons, French
FF8	French West Africa
FG8	Guadeloupe
FI8	French Indo-China
FK8	New Caledonia
FL8	Somaliland, French
FM8	Martinique
FN	French India
FO8	French Oceania (e.g., Tahiti)
FP8	St. Pierre & Miquelon Islands
FQ8	French Equatorial Africa
FR8	Reunion Island
FU8	New Hebrides
FY8	Guiana, French & Inini
G	England
GC	Channel Islands
GD	Isle of Man
GI	Ireland, Northern
GM	Scotland
GW	Wales
HA	Hungary
HB	Switzerland
HC	Ecuador
HE1	Liechtenstein
HH	Haiti
HI	Dominican Republic
HK	Colombia
HL	Korea
HP	Panama
HR	Honduras
HS	Siam
HV	Vatican City
HZ	Saudi Arabia (Hedjaz & Nejd)
I	Italy
I6	Eritrea
J	Japan
K	United States of America
KB6	Baker, Howland & American Phoenix Islands
KC6	Caroline Islands
KG4	Guantanamo Bay
KG6	Guam, Saipan, Tinian
KH6	Hawaiian Islands
KJ6	Johnston Islands
KL7	Alaska

KM6	Midway Islands
KP4	Puerto Rico
KP6	Palmyra Group, Jarvis Island
KR6	Ryukyu Islands (e.g., Okinawa)
KS6	American Samoa
KS4	Swan Island
KV4	Virgin Islands
KW6	Wake Island
KX6	Marshall Islands
KZ5	Canal Zone
LA	Norway
LI	Libya
LU	Argentina
LX	Luxembourg
LZ	Bulgaria
M1	San Marino
MB9	Austria
MD1	Cyrenaica
MD2	Tripolitania
MD3	Eritrea
MD4	Somalia
MD5	Suez Canal Zone
MD6	Iraq
MD7	Cyprus
MP4	Oman
OA	Peru
OE	Austria
OH	Finland
OK	Czechoslovakia
ON	Belgium
OQ	Belgian Congo
OX	Greenland
OY	Faeroes, The
OZ	Denmark
PA	Netherlands
PJ	Netherlands West Indies
PK1, 2, 3	Java
PK4	Sumatra
PK5	Borneo, Netherlands
PK6	Celebes & Molucca Islands
PK6	New Guinea, Netherlands
PX	Andorra
PY	Brazil
PZ	Guiana, Netherlands (e.g., Surinam)
SM	Sweden
SP	Poland
ST	Anglo-Egyptian Sudan
SU	Egypt
SV	Greece
SV	Crete
SV5	Dodecanese (e.g., Rhodes)
TA	Turkey
TF	Iceland
TG	Guatemala
TI	Costa Rica
TI	Cocos Island
UA1, 3, 4, 6	European Russian Socialist Federated Soviet Republic
UA9, Ø	Asiatic Russian S.F.S.R.
UB5	Ukraine
UC	White Russian Soviet Socialist Republic
UD6	Azerbaijan
UF6	Georgia
UG6	Armenia
UH8	Turkoman
UI8	Uzbek
UJ8	Tadzhik
UL7	Kazakh
UM8	Kirghiz
UN1	Karelo-Finnish Republic
UO5	Moldavia
UP	Lithuania
UQ	Latvia
UR	Estonia
VE	Canada
VK	Australia (including Tasmania)
VK1	Heard Island
VK1	MacQuarie Island
VK9	Papua Territory
VK9	New Guinea, Territory of
VK9	Norfolk Island

VO	Newfoundland & Labrador
VP1	British Honduras
VP2	Leeward & Windward Islands
VP3	British Guiana
VP4	Trinidad & Tobago
VP5	Jamaica & Cayman Islands
VP5	Turks & Caicos Islands
VP6	Barbados
VP7	Bahama Islands
VP8	Falkland Islands
VP8	South Georgia
VP8	South Orkney Islands
VP8	South Sandwich Islands
VP8	South Shetland Islands
VP9	Bermuda Islands
VQ1	Zanzibar
VQ2	Northern Rhodesia
VQ3	Tanganyika Territory
VQ4	Kenya
VQ5	Uganda
VQ6	Somaliland, British
VQ8	Mauritius & Chagos Islands
VQ9	Seychelles
VR1	Gilbert & Ellice Islands & Ocean Island
VR2	Fiji Islands
VR3	Fanning Island (Christmas Island)
VR4	Solomon Islands
VR5	Tonga (Friendly) Islands
VR6	Pitcairn Island
VS1	Strait Settlements
VS2	Federated Malay States
VS4	British North Borneo
VS5	Sarawak, Brunei
VS6	Hong Kong
VS7	Ceylon
VS8 (VU7)	Bahrein Islands
VS9	Maldive Islands
VU	India
VU4	Laccadive Islands
VU7	Bahrein Islands
W	United States of America
XE	Mexico
XZ	Burma
YA	Afghanistan
YI	Iraq
YK	Syria
YN	Nicaragua
YR	Roumania
YS	Salvador
YT, YU	Yugoslavia
YV	Venezuela
ZA	Albania
ZB1	Malta
ZB2	Gibraltar
ZC1	Trans-Jordan
ZC2	Cocos Islands
ZC3	Christmas Island
ZC4	Cyprus
ZC6	Palestine
ZD1	Sierra Leone
ZD2	Nigeria
ZD3	Gambia
ZD4	Togoland, Gold Coast
ZD6	Nyasaland
ZD7	St. Helena
ZD8	Ascension Island
ZD9	Tristan da Cunha & Gough Island
ZE1	Southern Rhodesia
ZK1	Cook Islands
ZK2	Niue
ZL	New Zealand
ZM	British Samoa
ZP	Paraguay
ZS1, 2, 4, 5, 6	Union of South Africa
ZS3	Southwest Africa
ZS7	Swaziland
ZS8	Basutoland
ZS9	Bechuanaland
3V8	Tunisia
4X4	Israel

Vacuum-Tube Data

For the convenience of the designer, the receiving-type tubes listed in this chapter are grouped by filament voltages and construction types (glass, metal, miniature, etc.). For example, all 6.3-volt metal tubes are listed in Table I, all lock-in base tubes are in Table III, all miniatures are in Table XI, and so on.

Transmitting tubes are divided into triodes and tetrodes-pentodes, then listed according to rated plate dissipation. This permits direct comparison of ratings of tubes in the same power classification.

For quick reference, all tubes are listed in numerical-alphabetical order in the index beginning on the following page.

Tube Ratings

Vacuum tubes are designed to be operated within definite maximum (and minimum) ratings. These ratings are the maximum safe operating voltages and currents for the electrodes, based on inherent limiting factors such as permissible cathode temperature, emission, and power dissipation in electrodes.

In the transmitting-tube tables, maximum ratings for electrode voltage, current and dissipation are given separately from the typical operating conditions for the recommended classes of operation. In the receiving-tube tables, because of space limitations, ratings and operating data are combined. Where only one set of operating conditions appears, the positive electrode voltages shown (plate, screen, etc.) are, in general, also the maximum rated voltages for those electrodes.

For certain air-cooled transmitting tubes, there are two sets of maximum values, one designated as CCS (Continuous Commercial Service) ratings, the other ICAS (Intermittent Commercial and Amateur Service) ratings. Continuous Commercial Service is defined as that type of service in which long tube life and reliability of performance under continuous operating conditions are the prime consideration. Intermittent Commercial and Amateur Service is defined to include the many applications where the transmitter design factors of minimum size, light weight, and maximum power output are more important than long tube life. ICAS ratings are considerably higher than CCS ratings. They permit the handling of greater power, and although such use involves some sacrifice in tube life, the period over which tubes will continue to give satisfactory performance in intermittent service can be extremely long.

Typical Operating Conditions

The typical operating conditions given for transmitting tubes represent, in general, maximum ICAS ratings where such ratings have been given by the manufacturer. They do not represent the *only* possible method of operation of a particular tube type. Other values of plate voltage, plate current, grid bias, etc., may be used so long as the maximum ratings for a particular voltage or current are not exceeded.

INDEX TO TUBE TABLES

BASE TYPE DESIGNATIONS

The type of base used on each tube listed in the tables is indicated in the base column by a letter whose meaning is as follows:

A = Acorn
B = Glass-button miniature
B_S = Glass-button subminiature
J = Jumbo
L = Lock-in
M = Medium
N = None or special type
O = Octal
S = Small
W = Wafer

INDEX TO VACUUM-TUBE TYPES

For convenience in locating data on specific tube types the index below lists all tubes in numerical-alphabetical order, showing the page number where individual tubes may be found in the classified-data section (pages 565–605) and the identifying base-diagram number in the base-diagram section (pages 558–564).

Type	Page	Base
12E5GT	574	6Q
12F5GT	574	5M
12G7G	574	7V
12H6	574	7Q
12J5GT	574	6Q
12J7GT	574	7R
12K7GT	574	7R
12K8	574	8K
12KP4	585	12D
12L8GT	574	8BU
12LP4	585	12D
12Q7GT	574	7V
12S8GT	574	8CD
12SA7	574	8R
12SC7	574	8S
12SF5	574	6AB
12SF7	574	7AZ
12SG7	574	8BK
12SH7	574	8BK
12SJ7	574	8N
12SK7	574	8N
12SL7GT	574	8BD
12SN7GT	574	8BD
12SQ7	574	8Q
12SR7	574	8Q
12SW7	574	8Q
12SX7	574	8BD
12SY7	574	8R
12Z3	586	4G
12Z5	586	7L
14A4	574	5AC
14A5	574	6AA
14A7	574	8V
14AF7	574	8AC
14AF7	576	8AC
14B6	574	8W
14B8	574	8X
14C5	574	6AA
14C7	575	8V
14E6	575	8W
14E7	575	8AE
14F7	575	8AC
14F8	575	8BW
14H7	575	8V
14J7	575	8BL
14N7	575	8AC
14Q7	575	8AL
14R7	575	8AE
14S7	575	8BL
14V7	575	8V
14W7	575	8BJ
14Y4	586	5AB
14Z3	586	4G
15	572	5F
15AP4	585	12D
15E	589	T-4AF
16AP4	585	12D
17	582	3G
18	575	6B
19	572	6C
19BG6G	575	5BT
19J6	580	7BF
19T8	580	9E
20BP4	585	12D
20	576	4D
20J8GM	575	8H
21A7	575	8AR
22	576	4K
24-A	571	5E
24-G	589	2D
24XH	585	Fig. 1
25A6	575	7S
25A7GT	575	8F
25A7GT	586	8F
25AC5G	575	6Q
25B5	575	6D
25B6G	575	7S
25B8GT	575	8T
25BQ6GT	575	6AM
25C6G	575	7AC
25D8GT	575	8AF
25L6	575	7AC
25N6G	575	7W
25S	571	6M
25T	589	3G
25W4	586	4CG
25X6GT	586	7Q
25Y4GT	586	5AA
25Y5	586	6E
25Z3	586	4G
25Z4	586	5AA
25Z5	586	6E
25Z6	586	7Q
26	576	4D
26A6	580	7BK
26A7GT	575	8BU
26C6	580	7BT
26D6	580	7CH
27	571	5A
28D7	576	8BS
28Z5	586	5AB
30	572	4D
31	572	4D
32	572	4K
32L7GT	575	8Z
32L7GT	586	8Z
33	572	5K
34	572	4M
35/51	571	5E
35A5	575	6AA
35B5	580	7BZ
35C5	580	7CV
35L6G	575	7AC
35T	591	3G
35TG	591	2D
35W4	586	5BQ
35Y4	586	5AL
35Z3	586	4Z
35Z4GT	586	5AA
35Z5G	586	6AD
35Z6G	587	7Q
36	570	5E
37	570	5A
38	570	5F
39/44	570	5F
40	576	4D
40Z5GT	587	6AD
41	570	6B
42	570	6B
43	575	6B
45	571	4D
45Z3	587	5AM
45Z5GT	587	6AD
46	571	5C
47	571	5B
48	575	6A
49	572	5C
50	576	4D
50A5	575	6AA
50B5	580	7BZ
50C5	580	7CV
50C6GT	575	7AC
50L6GT	575	7AC
50T	593	2D
50X6	587	7AJ
50Y6GT	587	7Q
50Y7GT	587	8AN
50Z6G	587	7Q
50Z7G	587	8AN
51	571	5E
52	570	5C
53	571	7B
53A	590	T-4B
55	571	6G
56	571	5A
56AS	570	5A
57	571	6F
57AS	570	6F
58	571	6F
58AS	570	6F
59	571	7A
70A7GT	575	8AB
70A7GT	587	8AB
70L7GT	575	8AA
70L7GT	587	8AA
71-A	577	4D
72	587	4P
73	587	4Y
75	570	6G
75TH	593	2D
75TL	593	2D
76	570	5A
77	570	6F
78	570	6F
79	570	6H
80	587	4C
81	587	4B
82	587	4C
83	587	4C
83-V	587	4AD
84/6Z4	587	5D
85	570	6G
85AS	570	6G
89	570	6F
99	577	4D
100TH	594	2D
100TL	594	2D
111H	593	2D
112-A	577	4D
117L7GT	575	8AO
117L7GT	587	8AO
117M7GT	575	8AO
117M7GT	587	8AO
117N7GT	575	8AV
117N7GT	587	8AV
117P7GT	575	8AV
117P7GT	587	8AV
117Z3	587	4BR
117Z4GT	587	5AA
117Z6GT	587	7Q
128AS	583	5A
150T	596	2N
152TH	596	4BC
152TL	596	4BC
182-B	577	4D
183	577	4D
203-A	594	4E
203-H	594	3N
204-A	598	T-1A
205D	588	4D
211	594	4E
212-E	598	T-2A
217A	587	4AT
217C	587	4AT
227A	594	T-4B
241B	598	T-2AA
242A	593	4E
242B	594	4E
242C	595	4E
249B	587	Fig. 53
250TH	597	2N
250TL	597	2N
254A	600	T-4C
254B	601	T-4C
261A	595	4E
270A	598	T-1A
276A	595	4E
282A	602	T-4C
284B	595	3N
284D	593	4E
295A	595	4E
300T	598	2N
303A	594	4E
304A	598	T-1A
304B	591	2D
304TH	598	4BC
304TL	598	4BC
305A	602	T-4CE
306A	600	T-5CB
307A	600	T-5C
308B	598	T-2A
310	589	4D
311	594	4E
311CH	595	Fig. 57
312A	602	T-6C
312E	598	T-2AA
316A	590	—
327A	594	T-4AD
327B	593	T-4AD
342B	594	4E
356A	592	T-4BD
361A	595	4E
376A	595	4E
410R	604	Fig. 58
482B	577	4D
483	577	4D
485	577	5A
527	598	T-4B
559	577	Fig. 18
703A	589	—
705A	587	T-3AA
707B	604	Fig. 61
715B	601	—
717A	568	8BK
723AB	604	Fig. 60
756	591	4D
800	590	2D
801	589	4D
801A	589	4D
802	599	6BM
803	603	5J
804	602	T-5C
805	596	3N
806	597	2N
807	600	5AW
808	592	2D
809	590	3G
810	596	2N
811	592	3G
811A	592	3G
812	592	3G
812A	592	3G
812H	593	3G
813	603	5BA
814	602	T-5D
815	600	8BY
816	587	4P
822	597	3N
822S	597	2N
826	592	7BO
828	602	5J
829	601	7BP
829A	601	7BP
829B	601	7BP
830	591	4D
830B	592	3G
831	598	T-1AA
832	600	7BP
832A	600	7BP
833A	598	T-1AB
834	592	2D
835	594	4E
836	587	4P
837	599	6BM
838	595	4E
840	572	5J
841	589	4D
841A	592	3G
841SW	592	3G
843	589	5A
844	600	5AW
849	598	T-1A
850	603	T-3B
852	595	2D
860	603	T-4CB
861	603	T-1B
864	577	4D
865	600	T-4C
866	587	4P
866A	587	4P
866B	587	4P
866Jr	587	4B
871	587	4P
872	587	4AT
872A	587	4AT
874	582	4S
876	582	—
878	587	4P
879	587	4P
884	582	6Q
885	582	5A
886	583	—
902	585	Fig. 1
903	585	6AL
904	585	Fig. 3
905	585	Fig. 6
906P1	583	7AN
907	585	Fig. 6
908	585	7AN
908A	585	7CE
909	585	Fig. 6
910	585	7AN
911	585	7AN
912	585	Fig. 8
913	585	Fig. 1
914	585	Fig. 12
930B	592	3G
938	595	4E
950	572	5K
951	571	4M
954	577	5BB
955	577	5BC
955	588	5BC
956	577	5BB
957	577	5BD
958	577	5BD
958A	577	5BD
958A	588	5BD
959	577	5BE
967	583	3G
975A	587	4AT
991	583	—
1003	587	4R
1005	587	5AQ
1006	587	4C
1201	577	8BN
1203	577	4AH
1204	577	8BO
1206	569	8BV
1221	570	6F
1223	568	7R
1229	572	4K
1230	572	4D
1231	569	8V
1232	569	8V
1247	581	—
1265	583	4AJ
1266	583	4AJ
1267	583	4V
1273	569	8V
1274	587	6S
1275	587	4C
1276	577	4D
1280	576	8V
1284	576	8V
1291	573	7BE
1293	573	4AA
1294	573	4AH
1299	573	6BB
1602	589	4D
1603	570	6F
1608	589	4D
1609	577	5B
1610	599	T-5CA
1611	566	7S
1612	566	7T
1613	599	7S
1614	600	7AC
1616	587	4P
1619	600	T-9H
1620	566	7R
1621	566	7S
1622	566	7AC
1623	590	3G
1624	601	T-5DC
1625	601	5AZ
1626	588	6Q
1627	596	2N
1628	590	T-4BB
1629	576	6RA
1631	576	7AC
1632	576	7AC
1633	576	8BD
1634	576	8S
1635	568	8B
1641	587	T-4AG
1642	569	7BH
1644	576	Fig. 7
1654	587	2Z
1800	585	6AL
1801	585	Fig. 13
1802P1	584	11A
1803P4	585	6AL
1804P4	584	6AL
1805P1	584	11A
1806P1	583	11A
1809P1	585	8BR
1811P1	584	6AZ
1851	566	7R
1852	565	8N
1853	565	8N
2001	585	4AA
2002	585	Fig. 1
2005	585	Fig. 1
2050	583	8BA
2051	583	8BA
24XH	585	Fig. 1
2523N/128AS	583	5A
5514	592	4BO
5516	600	7CL
5517	587	5BU
5556	588	4D
5562	601	Fig. 54
5590	580	7BD
5591	580	7BD
5618	599	7CU
5618	599	7CU
5633	582	—
5634	582	—
5637	581	—
5638	582	—
5640	582	—
5641	582	—
5642	582	—
5645	582	—
5648	595	—
5651	583	5BO
5654	580	7BD
5679	569	7CX
5687	580	9H
5691	568	8BD
5692	568	8BD
5693	566	8N
5722	580	5CB
5763	599	9K
5825	587	4P
7000	568	7R
7193	588	4AM
7700	570	6F
8000	597	2N
8001	602	7BM
8003	595	3N
8005	594	3G
8008	587	Fig. 11
8010-R	591	—
8012	590	T-4BB
8013-A	587	4P
8016	587	4AC
8020	587	4P
8025	590	4AQ
9001	580	7PM
9002	580	7TM
9002	588	7TM
9003	580	7PM
9004	577	4BJ
9005	577	5BG
9006	580	6BH
AT-340	603	5BK
BA	586	4J
BH	586	4J
BR	586	4H
CE220	586	4P
CK501	581	—
CK502	581	—
CK503	581	—
CK504	581	—
CK505	581	—
CK506	581	—
CK507	581	—
CK509	581	—
CK510	581	—
CK512	581	—
CK515BX	581	—
CK520AX	581	—
CK521AX	581	—
CK522AX	581	—
CK523AX	581	—
CK524AX	581	—
CK525AX	581	—
CK526AX	581	—
CK527AX	581	—
CK529AX	581	—
CK551AXA	581	—
CK553AXA	581	—
CK556AX	581	—
CK568AX	581	—
CK569AX	581	—
CK605CX	581	—
CK606BX	581	—
CK608CX	581	—
CK619CX	581	—
CK624CX	581	—
CK650AX	581	—
CK1005	587	5AQ
CK1006	587	4C
CK1007	587	T-9G
CK1009	587	—
CK5672	581	—
DR3B27	586	4B
DR123C	595	Fig. 26
DR200	596	2N
EF50	577	9C
F123A	595	Fig. 26
F127A	597	Fig. 26
G84	586	4B
GL2C44	577	Fig. 17
GL2C44	588	Fig. 17
GL5C24	597	Fig. 26
GL5D24	603	5BK
GL146	595	T-4BG
GL152	596	T-4BG
GL159	598	T-4BG
GL169	598	T-4BG
GL446A	577	Fig. 19
GL446A	588	Fig. 19
GL446B	588	Fig. 19
GL446B	577	Fig. 19
GL464A	588	Fig. 17
GL464A	577	Fig. 17
GL559	577	Fig. 18
GL592	597	Fig. 52
GL8012A	590	T-4BB
HD203A	596	3N
HF60	593	2D
HF75	593	2D
HF100	593	2D
HF120	594	4F
HF125	594	—
HF130	595	—
HF140	594	4F
HF150	595	—
HF175	595	T-3AC
HF200	596	2N
HF250	596	2N
HF300	597	2N
HK24	590	3G
HK54	591	2D
HK57	601	5BK
HK154	591	2D
HK158	591	2D
HK252L	596	4BC
HK253	587	4AT
HK254	594	2N
HK257	602	7BM
HK257B	602	7BM
HK304L	598	4BC
HK354	596	2N
HK354C	596	2N
HK354D	596	2N
HK354E	596	2N

Type	Page	Base
HK354F....	596	2N
HK454H ...	598	2N
HK454L....	598	2N
HK654.....	598	2N
HV12......	597	3N
HV18......	596	2N
HV27......	597	3N
HY6J5GTX	588	6Q
HY6L6GTX	600	7AC
HY6V6GTX	599	7AC
HY24	588	4D
HY25......	590	3G
HY30Z.....	590	4BO
HY31Z.....	590	T-4D
HY40......	590	3G
HY40Z.....	591	3G
HY51A.....	592	3G
HY51B.....	592	3G
HY51Z.....	592	4BO
HY57......	591	3G
HY60......	599	5AW
HY61......	600	5AW
HY63......	599	T-8DB
HY65......	600	T-8DB
HY67......	602	T-5DB
HY69......	601	T-5D
HY75......	589	2T
HY75A.....	589	2T
HY113.....	581	5K
HY114B....	588	2T
HY115.....	581	5K
HY123.....	581	5K
HY125.....	581	5K
HY145.....	581	5K
HY155.....	581	5K
HY615.....	588	T-8AG
HY801A....	589	4D
HY866jr....	587	4P
HY1231Z...	590	T-4D
HY1269....	601	T-5DB
HYE1148...	588	T-8AG
KY21......	583	—
KY866.....	583	Fig. 8

Type	Page	Base
M54.......	581	—
M64.......	581	—
M74.......	581	—
NU-2C35...	577	Fig. 38
PE340	603	5BK
QK159.....	604	Fig. 63
RK2J22....	605	—
RK2J23....	605	—
RK2J24....	605	—
RK2J25....	605	—
RK2J26....	605	—
RK2J27....	605	—
RK2J28....	605	—
RK2J29....	605	—
RK2J30....	605	—
RK2J31....	605	—
RK2J32....	605	—
RK2J33....	605	—
RK2J34....	605	—
RK2J36....	605	—
RK2J38....	605	—
RK2J39....	605	—
RK2J48....	605	—
RK2J49....	605	—
RK2J50....	605	—
RK2J54....	605	—
RK2J55....	605	—
RK2J56....	605	—
RK2J58....	605	—
RK2J61A...	605	—
RK2J62A...	605	—
RK2J66....	605	—
RK2J67....	605	—
RK2J68....	605	—
RK2J69....	605	—
RK4J31....	605	—
RK4J32....	605	—
RK4J33....	605	—
RK4J34....	605	—
RK4J35....	605	—
RK4J36....	605	—
RK4J37....	605	—
RK4J38....	605	—

Type	Page	Base
RK4J39....	605	—
RK4J40....	605	—
RK4J41....	605	—
RK4J43....	605	—
RK4J44....	605	—
RK4J53....	605	—
RK4J54....	605	—
RK4J55....	605	—
RK4J56....	605	—
RK4J57....	605	—
RK4J58....	605	—
RK4J59....	605	—
RK725A....	605	—
RK10......	589	4D
RK11......	590	3G
RK12......	590	3G
RK15......	571	4D
RK16......	571	5A
RK17......	571	5F
RK18......	590	3G
RK19......	587	4AT
RK20......	601	T-5C
RK20A.....	601	T-5C
RK21......	587	4P
RK22......	587	T-4AG
RK23......	599	6BM
RK24......	572	4D
RK24......	588	4D
RK25......	599	6BM
RK25B.....	599	6BM
RK28......	603	5J
RK28A.....	603	5J
RK30......	590	2D
RK31......	590	3G
RK32......	591	2D
RK33......	588	7BH
RK34......	588	T-7DC
RK35......	591	2D
RK36......	594	2D
RK37......	591	2D
RK38......	594	2D
RK39......	600	5AW
RK41......	600	5AW

Type	Page	Base
RK42......	573	4D
RK43......	573	6C
RK44......	599	6BM
RK46......	601	T-5C
RK47......	602	T-5D
RK48......	603	T-5D
RK48A.....	602	T-5D
RK49......	600	6A
RK51......	592	3G
RK52......	592	3G
RK56......	599	5AW
RK57......	595	3N
RK58......	594	3N
RK59......	589	T-4D
RK60......	587	T-4AG
RK61......	581	—
RK62......	583	4D
RK63......	597	2N
RK63A.....	597	2N
RK64	599	5AW
RK65......	603	T-3BC
RK66......	601	T-5C
RK75......	600	T-5C
RK100.....	589	6A
RK705A....	587	T-3AA
RK866.....	587	4P
RM208.....	583	—
RM209.....	583	—
SD917A....	581	—
SD828A....	582	—
SD828E....	582	—
SN944.....	582	—
SN946	582	—
SN947D....	582	—
SN948C....	582	—
SN953.....	582	—
SN954.....	582	—
SN955B....	582	—
SN956B....	582	—
SN957A....	582	—
SN1006	582	—
SN1007B...	582	—
T20.......	589	3G

Type	Page	Base
T21.......	600	6A
T40.......	591	3G
T55.......	592	3G
T60.......	592	2D
T100......	593	2D
T125......	595	2N
T200......	597	2N
T300......	597	—
T814......	597	3N
T822......	597	3N
TB35	601	Fig. 54
TUF20.....	589	2T
TW75......	593	2D
TW150.....	596	2N
TZ20......	589	3G
TZ40	591	3G
UE100.....	593	2D
UE468.....	596	Fig. 57
UH35......	593	3G
UH50......	591	2D
UH51......	591	2D
V70.......	593	3N
V70A......	593	3N
V70B......	593	3G
V70C......	593	3G
V70D......	594	3G
VR75......	583	4AJ
VR90......	583	4AJ
VR105	583	4AJ
VR150	583	4AJ
VT52......	577	4D
VT127A....	594	T-4B
WE304A ...	591	2D
X6030.....	577	Fig. 4
XXB	578	Fig. 9
XXD......	576	8AC
XXL......	569	5AC
XXFM.....	578	8BZ
Z225......	587	4P
Z668......	604	—
ZB60......	593	2D
ZB120	593	4E

VACUUM-TUBE BASE DIAGRAMS

The diagrams on the following pages show standard socket connections corresponding to the base designations given in the column headed "Socket Connections" in the classified tube-data tables. Bottom views are shown throughout. Terminal designations are as follows:

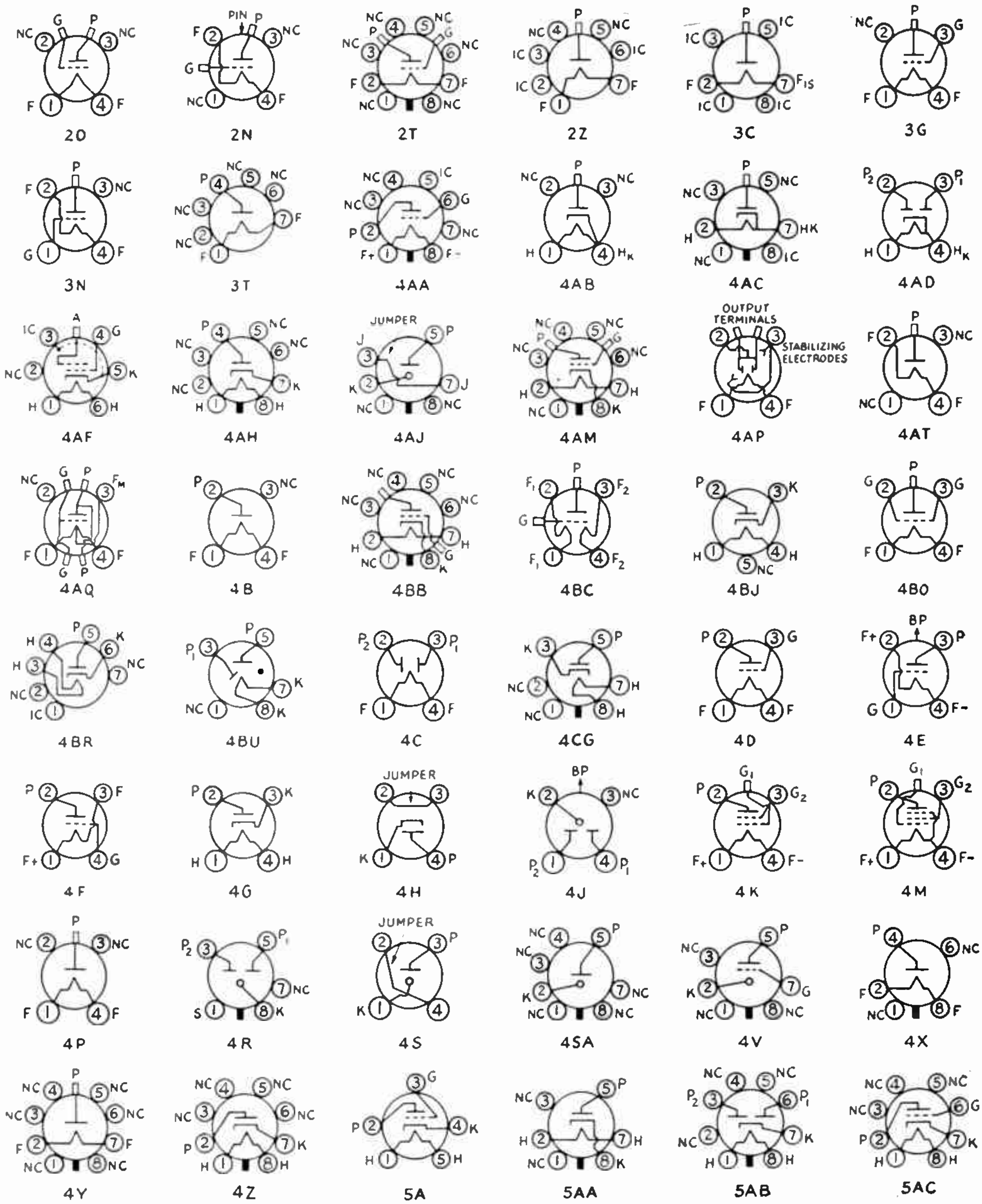
A = Anode	F = Filament	IS = Internal Shield	P_{BF} = Beam-Forming Plates	repeller
B = Beam	G = Grid	K = Cathode	RC = Ray-Control Electrode	S = Shell
BP = Bayonet Pin	H = Heater	NC = No Connection	Ref = Reflector or	TA = Target
BS = Base sleeve	IC = Internal Connection	P = Plate (Anode)		● = Gas-Type Tube
D = Deflecting Plate		P_1 = Starter-Anode		U = Unit

Alphabetical subscripts D, P, T and HX indicate, respectively, diode unit, pentode unit, triode unit or hexode unit in multi unit types. Subscript M, T or CT indicates filament or heater tap.

Generally when the No. 1 pin of a metal-type tube in Table I, with the exception of all triodes, is shown connected to the shell, the No. 1 pin in the glass (G or GT) equivalent is connected to an internal shield.

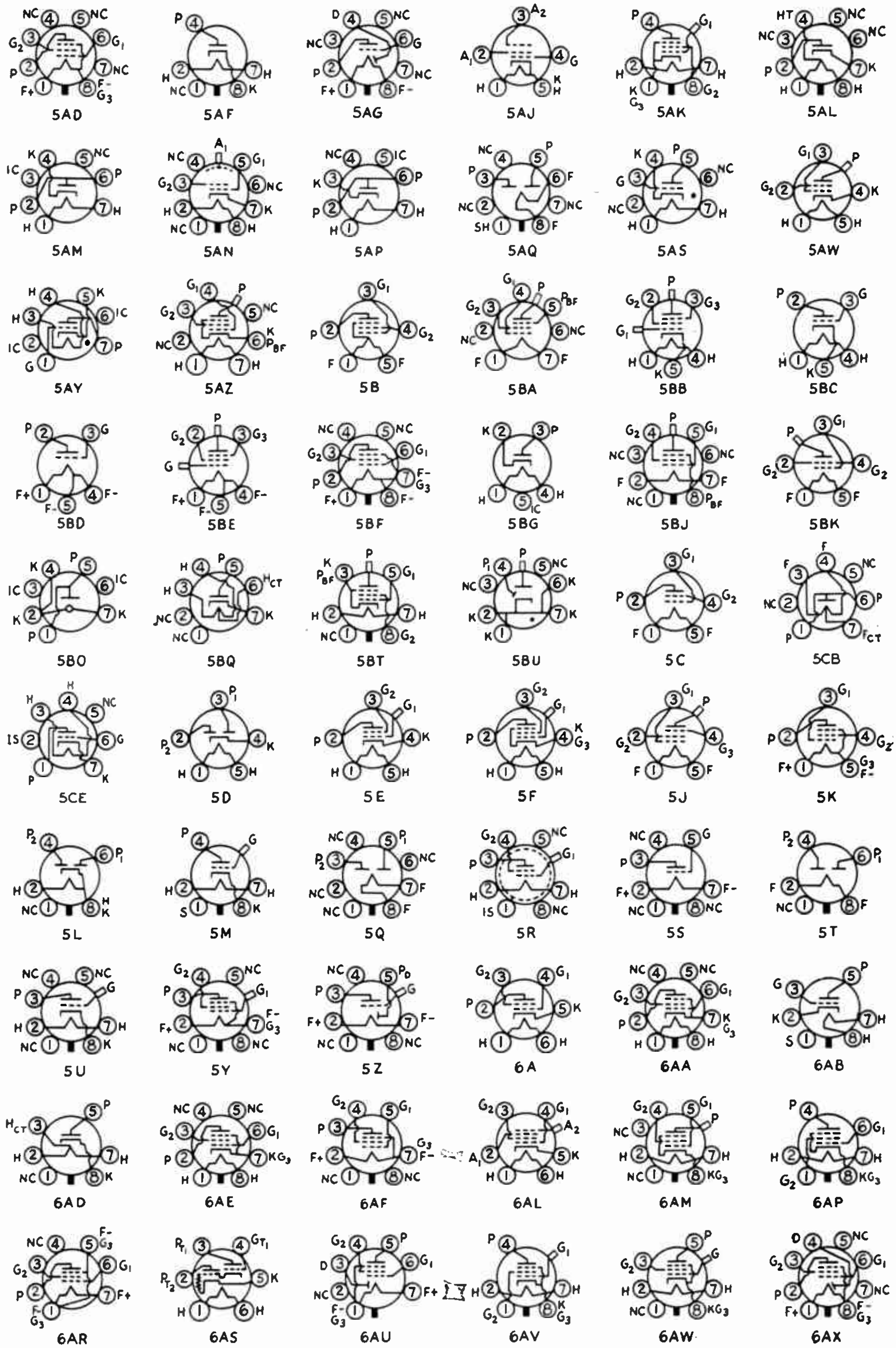
R.M.A. TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are shown above.



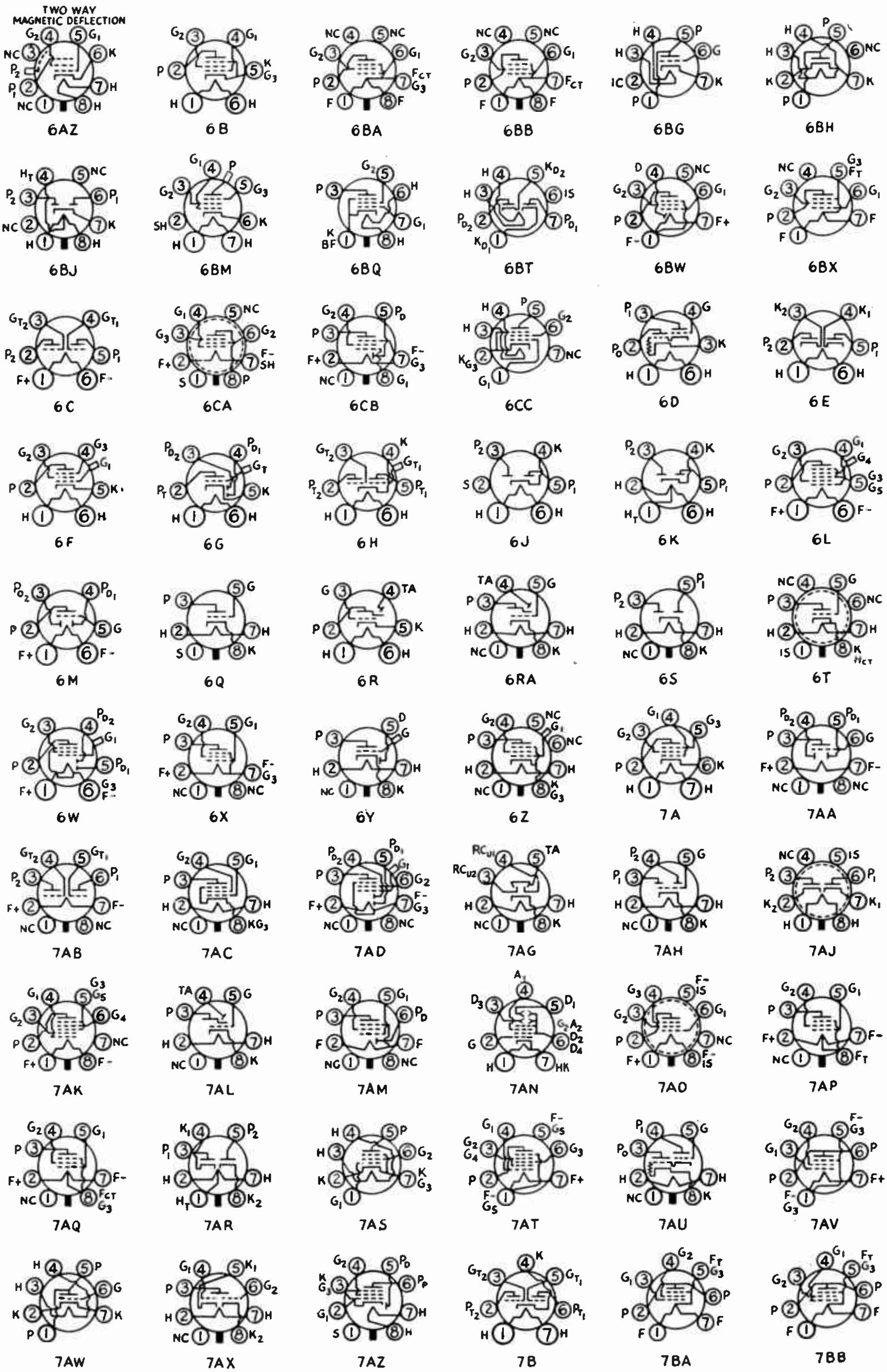
R.M.A. TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page 558.



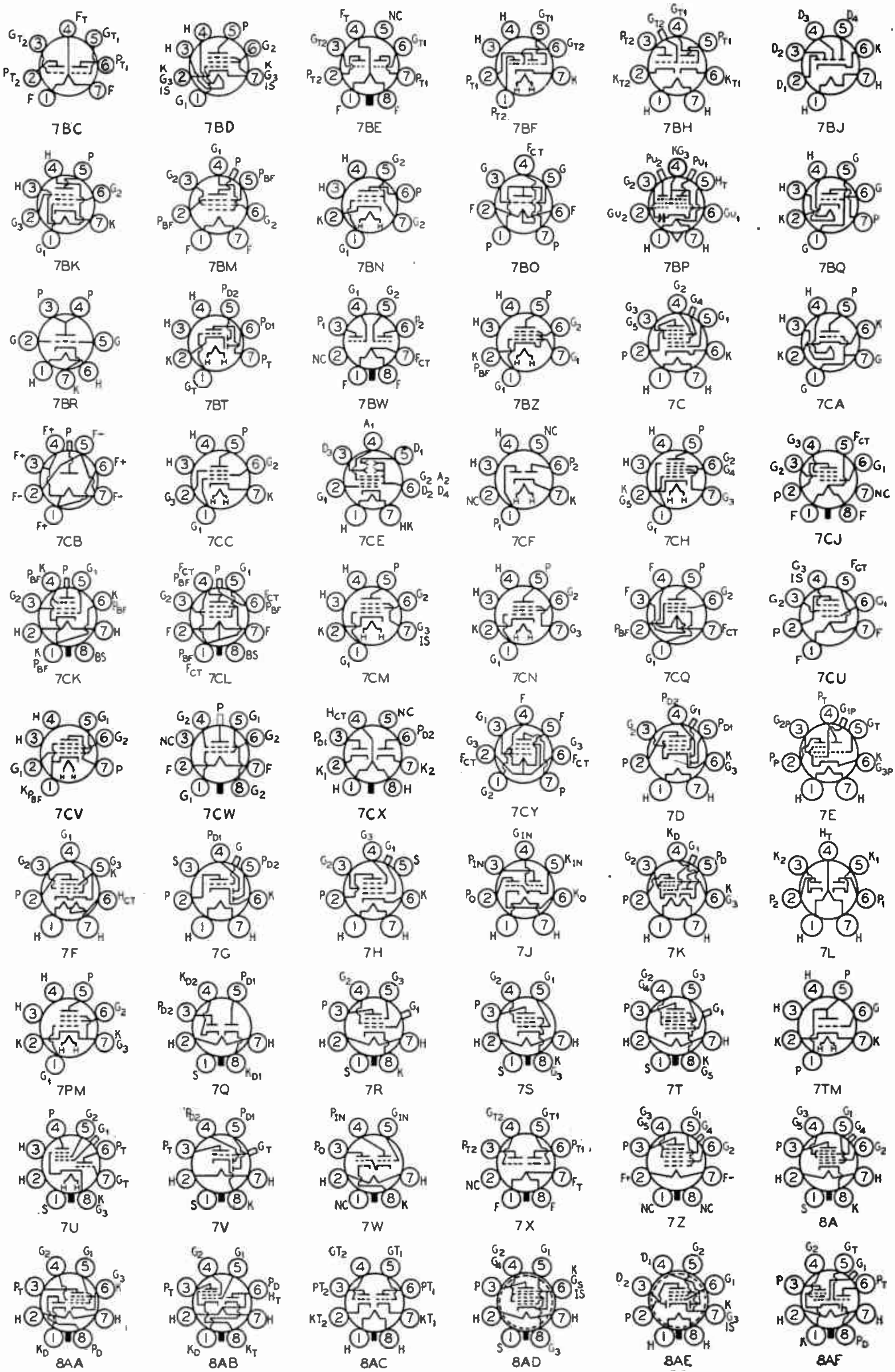
R.M.A. TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page 558.



R.M.A. TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page 558.



7BC 7BD 7BE 7BF 7BH 7BJ

7BK 7BM 7BN 7BO 7BP 7BQ

7BR 7BT 7BW 7BZ 7C 7CA

7CB 7CC 7CE 7CF 7CH 7CJ

7CK 7CL 7CM 7CN 7CQ 7CU

7CV 7CW 7CX 7CY 7D 7E

7F 7G 7H 7J 7K 7L

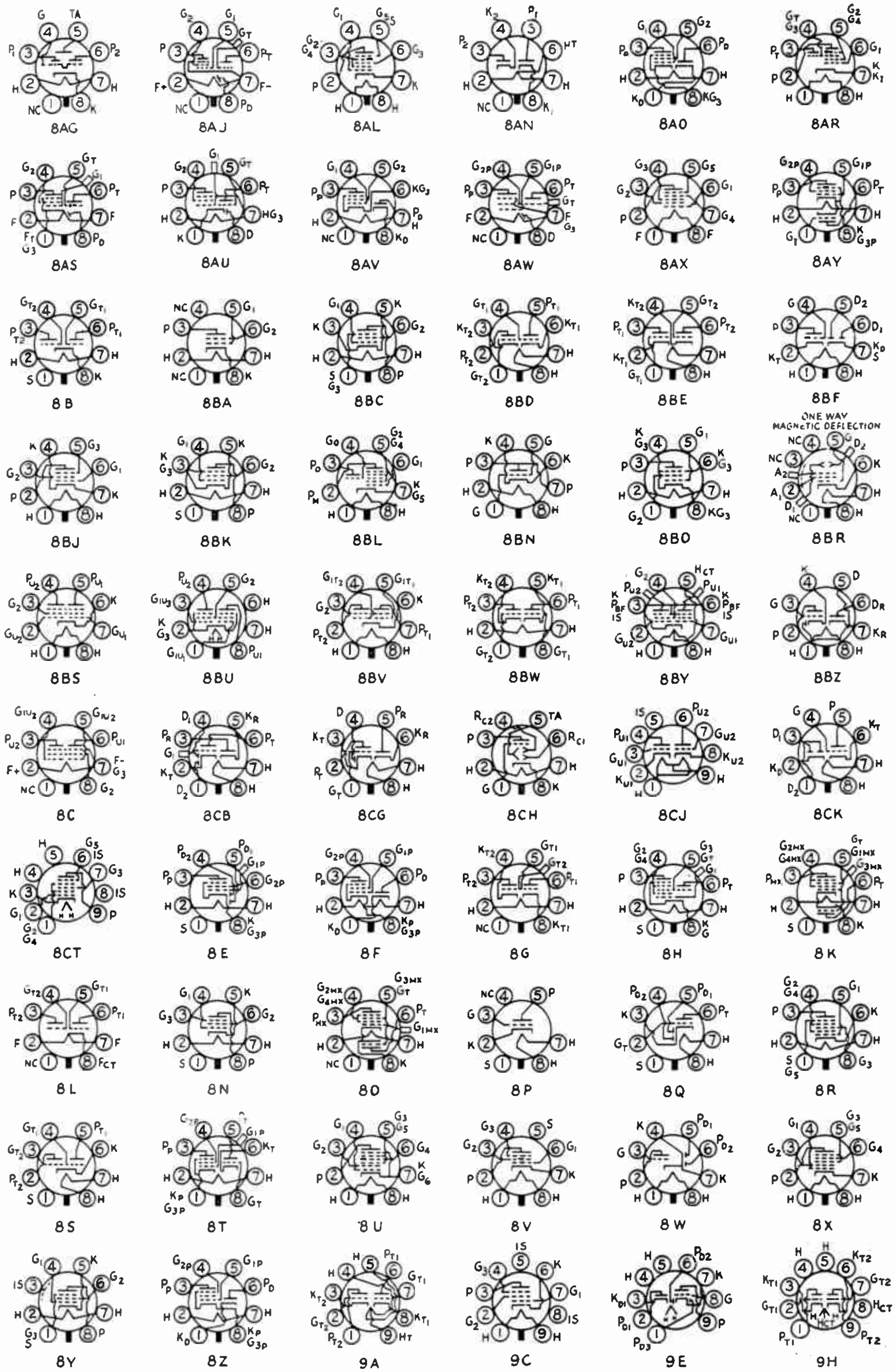
7PM 7Q 7R 7S 7T 7TM

7U 7V 7W 7X 7Z 8A

8AA 8AB 8AC 8AD 8AE 8AF

R.M.A. TUBE BASE DIAGRAMS

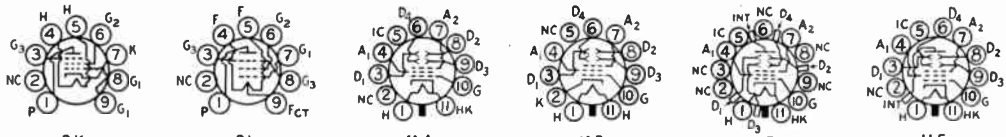
Bottom views are shown. Terminal designations on sockets are given on page 558.



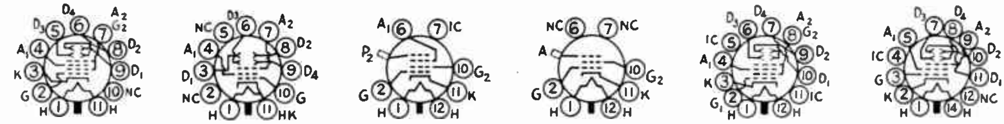
8AG 8AJ 8AL 8AN 8AO 8AR
8AS 8AU 8AV 8AW 8AX 8AY
8B 8BA 8BC 8BD 8BE 8BF
8BJ 8BK 8BL 8BN 8BO 8BR
8BS 8BU 8BV 8BW 8BY 8BZ
8C 8CB 8CG 8CH 8CJ 8CK
8CT 8E 8F 8G 8H 8K
8L 8N 8O 8P 8Q 8R
8S 8T 8U 8V 8W 8X
8Y 8Z 9A 9C 9E 9H

R.M.A. TUBE BASE DIAGRAMS

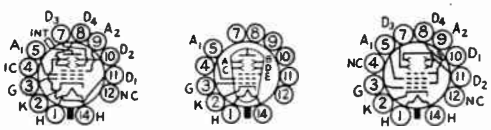
Bottom views are shown. Terminal designations on sockets are given on page 558.



9K 9L 11A 11B 11E 11F



11M 11N 12C 12D 12E 14A



14B 14F 14G

SUPPLEMENTARY BASE DIAGRAMS

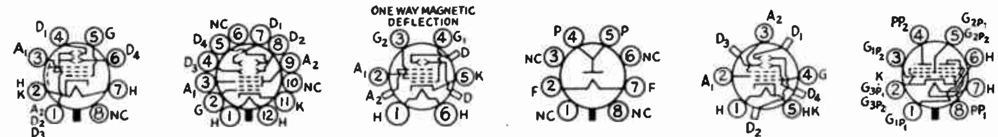


FIG. 1 FIG. 2 FIG. 3 FIG. 4 FIG. 6 FIG. 7

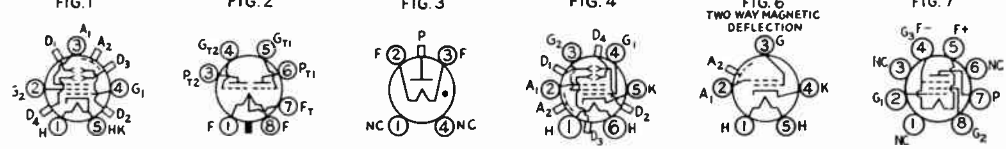


FIG. 8 FIG. 9 FIG. 11 FIG. 12 FIG. 13 Fig. 14

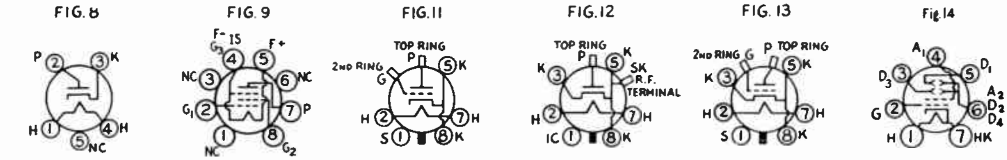


FIG. 15 FIG. 16 FIG. 17 FIG. 18 FIG. 19 FIG. 20

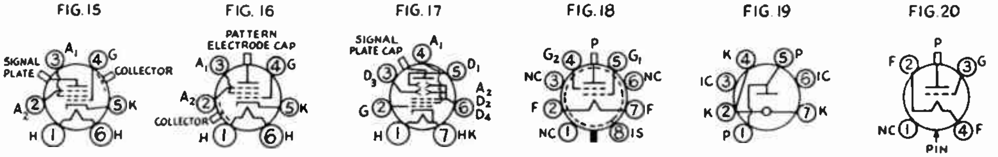


FIG. 21 FIG. 22 FIG. 23 FIG. 24 FIG. 25 FIG. 26

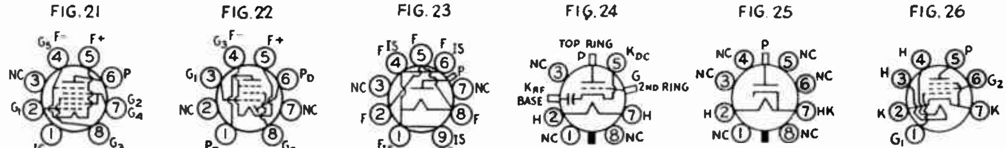


FIG. 27 FIG. 28 FIG. 29 FIG. 30 FIG. 31 FIG. 32

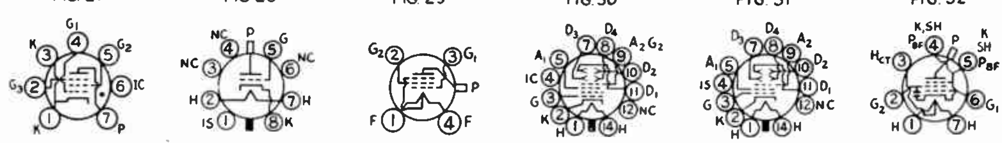


FIG. 33 FIG. 38 FIG. 40 FIG. 47 FIG. 49 FIG. 50

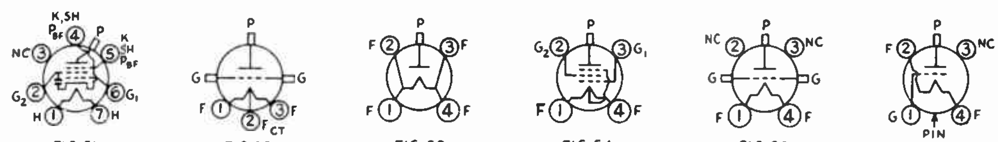


FIG. 51 FIG. 52 FIG. 53 FIG. 54 FIG. 56 FIG. 57

SUPPLEMENTARY BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page 558.

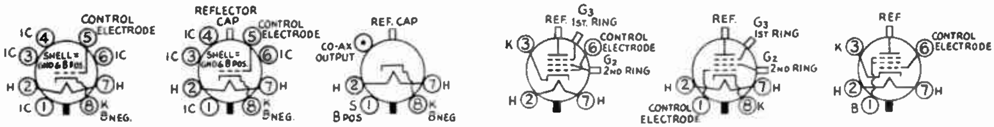
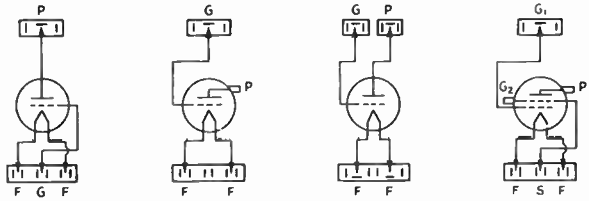
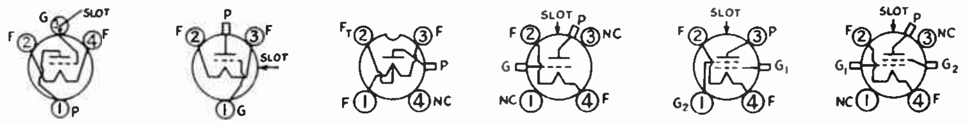


FIG. 58 FIG. 59 FIG. 60 FIG. 61 FIG. 62 FIG. 63

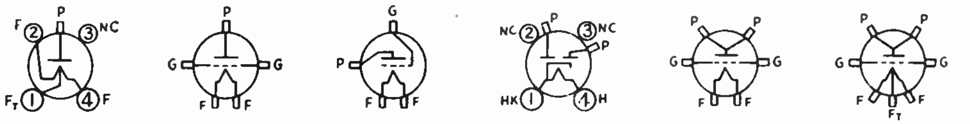
SUPPLEMENTARY "T"-GROUP BASE DIAGRAMS



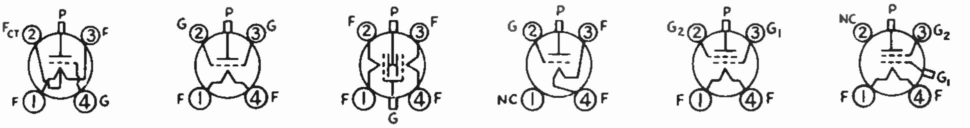
T-1A T-1AA T-1AB T-1B



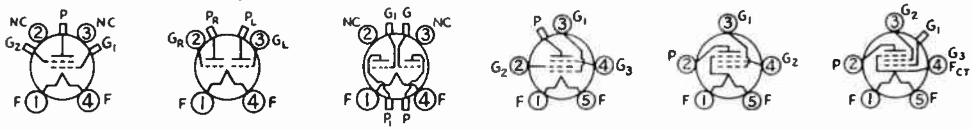
T-2A T-2AA T-3AA T-3AC T-3B T-3BC



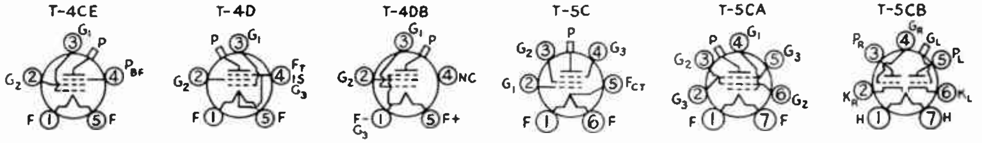
T-4A T-4AD T-4AF T-4AG T-4B T-4BB



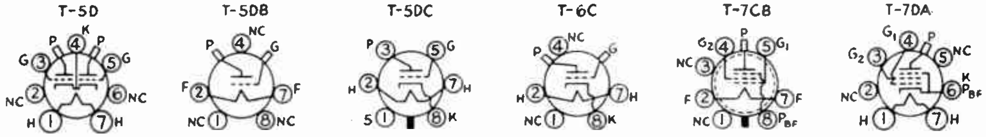
T-4BD T-4BE T-4BF T-4BG T-4C T-4CB



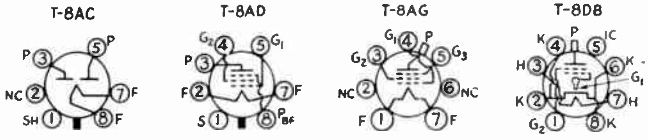
T-4CE T-4D T-4DB T-5C T-5CA T-5CB



T-5D T-5DB T-5DC T-6C T-7CB T-7DA



T-7DC T-8AC T-8AD T-8AG T-8DB T-9E



T-9G T-9H T-9I T-9J

TABLE I—METAL RECEIVING TUBES

Characteristics given in this table apply to all tubes having type numbers shown, including metal tubes, glass tubes with "G" suffix, and bantam tubes with "GT" suffix. For "G" and "GT" tubes not listed (not having metal counterparts), see Tables II, VII, VIII and IX.

Type	Name	Socket Connections	Fil. or Heater		Capacitance μμfd.			Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
			Volts	Amp.	In	Out	Plate-Grid												
6A8	Pentagrid Convertor	8A	6.3	0.3	—	—	—	Osc.-Mixer	250	– 3.0	100	3.2	3.3	Anode-grid (No. 2) 250 volts max. thru 20,000 ohms					6A8
6AB7 1853	Television Amp. Pentode	8N	6.3	0.45	8	5	0.015	Class-A Amp.	300	– 3.0	200	3.2	12.5	700000	5000	3500	—	—	6AB7 1853
6AC7 1852	Television Amp. Pentode	8N	6.3	0.45	11	5	0.015	Class-A Amp.	300	160*	150	2.5	10	1000000	9000	6750	—	—	6AC7 1852
6AG7	Sharp Cut-off Pentode	8Y	6.3	0.65	13	7.5	0.06	Class-A_1 Amp.	300	– 3.0	150	7/9	30/30.5	130000	11000	—	10000	3.0	6AG7
6AJ7	Sharp Cut-off Pentode	8N	6.3	0.45	—	—	—	Class-A Amp.	300	160*	300	2.5	10	1000000	9000	—	—	—	6AJ7
6AK7	Pentode Power Amp.	8Y	6.3	0.65	13	7.5	0.06	Class-A Amp.	300	– 3	150	7	30	130000	11000	—	10000	3.0	6AK7
6B8	Duplex-Diode Pentode	8E	6.3	0.3	6	9	0.005	Class-A Amp.	250	– 3.0	125	2.3	9.0	650000	1125	730	—	—	6B8
6C5	Triode	6Q	6.3	0.3	3	11	2	Class-A Amp.	250	– 8.0	—	—	8.0	10000	2000	20	—	—	6C5
								Bias Detector	250	–17.0	—	—	Plate current adjusted to 0.2 ma. with no signal						
6F5	High-μ Triode	5M	6.3	0.3	5.5	4	2.3	Class-A Amp.	250	– 1.3	—	—	0.2	66000	1500	100	—	—	6F5
6F6	Pentode Power Amplifier	7S	6.3	0.7	6.5	13	0.2	Class-A_1 Pent.[5]	250 315	–16.5 –22.0	250 315	6.5 8.0	36[7] 42	80000 75000	2500 2650	200 200	7000 7000	3.2 5.0	6F6
								Class-A_1 Triode[1]	250	–20.0	—	—	34[7]	2600	2600	6.8	4000	0.85	
								Class-AB_2 Amp.[6] Class-AB_2 Amp.[6]	375 375	340* –26.0	250 250	8/18 5/19.5	54/77 34/82	Power output for 2 tubes at stated load, plate-to-plate			10000[8] 10000[8]	19.0 18.5	
								Class-AB_2 Amp.[1,6]	350 350	730* –38	— —	— —	50/61 48/92	— —	— —	— —	10000[8] 6000[8]	9 13	
6H6	Twin Diode	7Q	6.3	0.3	—	—	—	Rectifier		Max. a.c. voltage per plate = 150 r.m.s. Max. output current 8.0 ma. d.c.									6H6
6J5	Triode	6Q	6.3	0.3	3.4	3.6	3.4	Class-A Amp.	250	– 8.0	—	—	9	7700	2600	20	—	—	6J5
6J7	Sharp Cut-off Pentode	7R	6.3	0.3	7	12	0.005	R.F. Amp.	250	– 3.0	100	0.5	2.0	1.5 meg.	1225	1500	—	—	6J7
								Bias Detector	250	– 4.3	100	Cathode current 0.43 ma.			—	—	0.5 meg.	—	
6K7	Variable-μ Pentode	7R	6.3	0.3	7	12	0.005	R.F. Amp.	250	– 3.0	125	2.6	10.5	600000	1650	990	—	—	6K7
								Mixer	250	–10.0	100	—	—	—	Oscillator peak volts = 7.0				
6K8	Triode-Hexode	8K	6.3	0.3	—	—	—	Converter	250	– 3.0	100	6	2.5	Triode Plate (No. 2) 100 volts, 3.8 ma.					6K8
6L6	Beam Power Amplifier	7AC	6.3	0.9	10	12	0.4	Single Tube Class A_1	250 300	170* 220*	250 200	5.4/7.2 3.0/4.6	75/78 51/54.5	— —	— —	— —	2500 4500	6.5 6.5	6L6
								Single Tube Class A_1	250 350	–14.0 –18.0	250 250	5.0/7.3 2.5/7.0	72/79 54/66	22500 33000	6000 5200	— —	2500 4200	6.5 10.8	
								P.P. Class A_1[6]	270	125*	270	11/17	134/145	—	—	—	5000[8]	18.5	
								P.P. Class A_1[6]	250 270	–16.0 –17.5	250 270	10/16 11/17	120/140 134/155	24500 23500	5500 5700	— —	5000[8] 5000[8]	14.5 17.5	
								P.P. Class AB_1[3]	360	250*	270	5/17	88/100	Power output for 2 tubes. Load plate-to-plate			9000[8]	24.5	
								P.P. Class AB_1[6]	360	–22.5	270	5/15	88/132				6600[8]	26.5	
								P.P. Class AB_2[6]	360 360	–18.0 –22.5	225 270	3.5/11 5/16	78/142 88/205				6000[8] 3800[8]	31.0 47.0	
6L7	Pentagrid Mixer Amplifier	7T	6.3	0.3	—	—	—	R.F. Amp.	250	– 3.0	100	5.5	5.3	800000	1100	—	—	—	6L7
								Mixer	250	– 6.0	150	8.3	3.3	Over 1 meg.	Oscillator-grid (No. 3) voltage = –15				
6N7	Twin Triode	8B	6.3	0.8	—	—	—	Class-B Amp.	300	0	—	—	35/70	—	—	—	8000	10.0	6N7
6Q7	Duplex-Diode Triode	7V	6.3	0.3	5	3.8	1.4	Triode Amp.	250	– 3.0	—	—	1.1	58000	1200	70	—	—	6Q7
6R7	Duplex-Diode Triode	7V	6.3	0.3	4.8	3.8	2.4	Triode Amp.	250	– 9.0	—	—	9.5	8500	1900	16	10000	0.28	6R7
6S7	Remote Cut-off Pentode	7R	6.3	0.15	6.5	10.5	0.005	Class-A Amp.	250	– 3.0	100	2.0	8.5	1000000	1750	—	—	—	6S7
6SA7	Pentagrid Converter	8R[2]	6.3	0.3	—	—	—	Converter	250	0[3]	100	8.0	3.4	800000	Grid No. 1 resistor 20000 ohms				6SA7
B7Y	Pentagrid Converter	8R	6.3	0.3	9.6	9.2	—	Converter	100	– 1	100	10.2	3.6	500000	900	—	—	—	6SB7Y
								Converter	250	– 1	100	10	3.8	1000000	950	—	—	—	
					Osc. Section in 88–108 Mc. Serv.				250	22000[9]	12000[9]	12.6/12.5	6.8/6.5	—	—	—	—	—	
'7	Twin-Triode	8S	6.3	0.3	—	—	—	Class-A Amp.	250	– 2.0	—	—	2.0	53000	1325	70	—	—	6SC7
	High-μ Triode	6AB	6.3	0.3	4	3.6	2.4	Class-A Amp.	250	– 2.0	—	—	0.9	66000	1500	100	—	—	6SF5

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TABLE I—METAL RECEIVING TUBES—*Continued*

	Name	Socket Connections	Fil. or Heater		Capacitance μμfd.			Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
			Volts	Amp.	In	Out	Plate-Grid												
	e Variable-μ Pentode	7AZ	6.3	0.3	5.5	6	0.004	Class-A Amp.	250	– 1.0	100	3.3	12.4	700000	2050	—	—	—	6SF7
	mivariable-μ Pentode	8BK	6.3	0.3	8.5	7	0.003	H.F. Amp.	250	– 2.5	150	3.4	9.2	Over 1 meg.	4000	—	—	—	6SG7
	Sharp Cut-off Pentode	8BK	6.3	0.3	8.5	7	0.003	Class-A Amp.	250	– 1.0	150	4.1	10.8	900000	4900	—	—	—	6SH7
	Sharp Cut-off Pentode	8N	6.3	0.3	6	7	0.005	Class-A Amp.	250	– 3.0	100	0.8	3	1500000	1650	2500	—	—	6SJ7
K7	Variable-μ Pentode	8N	6.3	0.3	6	7	0.003	Class-A Amp.	250	– 3.0	100	2.4	9.2	800000	2000	1600	—	—	6SK7
6SQ7	Duplex-Diode Triode	8Q	6.3	0.3	3.2	3.0	1.6	Class-A Amp.	250	– 2.0	—	—	0.8	91000	1100	100	—	—	6SQ7
6SR7	Duplex-Diode Triode	8Q	6.3	0.3	3.6	2.8	2.40	Class-A Amp.	250	– 9.0	—	—	9.5	8500	1900	16	—	—	6SR7
6SS7	Variable-μ Pentode	8N	6.3	0.15	5.5	7.0	0.004	Class-A Amp.	250	– 3.0	100	2.0	9.0	1000000	1850	—	—	—	6SS7
6ST7	Duplex-Diode Triode	8Q	6.3	0.15	2.8	3	1.50	Class-A Amp.	250	– 9.0	—	—	9.5	8500	1900	16	—	—	6ST7
6SV7	Diode R.F. Pentode	7AZ	6.3	0.3	6.5	6	0.004	Class-A Amp.	250	– 1	150	2.8	7.5	800000	3400	—	—	—	6SV7
6SZ7	Duplex-Diode Triode	8Q	6.3	0.15	2.6	2.8	1.10	Class-A Amp.	250	– 3	—	—	1.0	58000	1200	70	—	—	6SZ7
6T7	Duplex-Diode Triode	7V	6.3	0.15	1.8	3.1	1.70	Class-A Amp.	250	– 3.0	—	—	1.2	62000	1050	65	—	—	6T7
6V6	Beam Power Amplifier	7AC	6.3	0.45	2.0	7.5	0.7	Class-A_1 Amp.[5]	250	–12.5	250	4.5/7.0	45/47	52000	4100	218	5000	4.5	6V6
								Class-AB_1 Amp.[6]	250	–15.0	250	5/13	70/79	60000	3750	—	10000[8]	10.0	
									285	–19.0	285	4/13.5	70/92	65000	3600	—	8000[8]	14.0	
1611	Pentode Power Amplifier	7S	6.3	0.7	—	—	—	Audio Amp.	Characteristics same as 6F6										1611
1612	Pentagrid Amplifier	7T	6.3	0.3	7.5	11	0.001	Class-A Amp.	250	– 3.0	100	6.5	5.3	600000	1100	880	—	—	1612
1620	Sharp Cut-off Pentode	7R	6.3	0.3	—	—	—	Class-A Amp.	Characteristics same as 6J7										1620
1621	Power Amplifier Pentode	7S	6.3	0.7	—	—	—	Class-AB_2 Amp.[6]	300	–30.0	300	6.5/13	38/69	—	—	—	4000[8]	5.0	1621
								Class-A_1 Amp.	330	500*	—	—	55/59	—	—	—	5000[8]	2.0	
1622	Beam Power Amplifier	7AC	6.3	0.9	—	—	—	Class-A_1 Amp.	300	–20.0	250	4/10.5	86/125	—	—	—	4000	10.0	1622
1851	Television Amp. Pentode	7R	6.3	0.45	11.5	5.2	0.02	Class-A Amp.	300	– 2.0	150	2.5	10	750000	9000	6750	—	—	1851
5693	Sharp Cut-off Pentode	8N	6.3	0.3	5.3	6.2	0.005	Class-A Amp.	250	– 3	100	0.85	3.0	1000000	1650	—	—	—	5693

* Cathode resistor—ohms.
[1] Screen tied to plate.
[2] For 6SA7GT use base diagram 8AD.
[3] Grid bias—2 volts if separate oscillator excitation is used.
[4] Also Type "6SJ7Y."
[5] Values are for single tube.
[6] Values are for two tubes in push-pull.
[7] Max.-signal value.
[8] Plate-to-plate value.
[9] Osc. grid leak—5crn res.

TABLE II—6.3-VOLT GLASS TUBES WITH OCTAL BASES

(For "G" and "GT"-Type Tubes Not Listed Here, See Equivalent Type in Table I; Characteristics and Connections Will Be Identical)

Type	Name	Socket Connections	Fil. or Heater		Capacitance μμfd.			Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
			Volts	Amp.	In	Out	Plate-Grid												
2C22	Triode	4AM	6.3	0.3	2.2	0.7	3.60	Class-A Amp.	300	–10.5	—	—	11	6600	3000	20	—	—	2C22
6A5G	Triode Power Amplifier	6T	6.3	1.0	—	—	—	Class-A Amp.[4]	250	–45.0	—	—	60	800		4.2	2500	3.75	6A5G
								P.P. Class AB[5]	325	–68.0	—	—	80	—	5250	—	3000[6]	15.0	
								P.P. Class AB[5]	325	850*			80	—		—	5000[6]	10.0	
6AB6G	Direct-Coupled Amplifier	7AU	6.3	0.5	—	—	—	Class-A Amp.	250	0	Input		5.0	40000	1800	72	8000	3.5	6AB6G
									250	0	Output		34						
6AC5G	High-μ Power-Amplifier Triode	6Q	6.3	0.4	—	—	—	P.P. Class B[6]	250	0	—	—	5.0	36700	3400	125	10000[6]	8.0	6AC5G
								Dyn.-Coupled	250	—	—	—	32				7000	3.7	
6AC6G	Direct-Coupled Amplifier	7AU	6.3	1.1	—	—	—	Class-A Amp.	180	0	Input		7.0	—	3000	54	4000	3.8	6AC6G
									180	0	Output		45						
6AD5G	High-μ Triode	6Q	6.3	0.3	4.1	3.9	3.3	Class-A Amp.	250	– 2.0	—	—	0.9	—	1500	100	—	—	6AD5G
6AD6G[10]	Electron-Ray Tube	7AG	6.3	0.15	—	—	—	Indicator	100		0 for 90°; —23 for 135°; 45 for 0°. Target current 1.5 ma.								6AD6G
6AD7G	Triode-Pentode	8AY	6.3	0.85	—	—	—	Triode Amp.	250	–25.0	—	—	4.0	19000	325	6.0	—	—	6AD7G
								Pentode Amp.	250	–16.5	250	6.5	34	80000	2500	—	7000	3.2	
6AE5G[10]	Triode Amplifier	6Q	6.3	0.3	—	—	—	Class-A Amp.	95	–15.0	—	—	7.0	3500	1200	4.2	—	—	6AE5G
6AE6GT[10]	Twin-Plate Triode with Single Grid	7AH	6.3	0.15	Remote cut-off			Class-A Amp.	250	– 1.5	—	—	6.5	25000	1000	25	—	—	6AE6GT
					Sharp cut-off			Class-A Amp.	250	– 1.5			4.5	35000	950	33			

TABLE II—6.3-VOLT GLASS TUBES WITH OCTAL BASES—*Continued*

Type	Name	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
6AE7GT[10]	Twin-Input Triode	7AX	6.3	0.5	—	—	—	Driver Amplifier	250	−13.5	—	—	5.0	9300	1500	14	—	—	6AE7GT
6AF5G	Triode	6Q	6.3	0.3	—	—	—	Class-A Amplifier	180	−18.0	—	—	7.0	—	1500	7.4	—	—	6AF5G
6AF7G	Twin Electron Ray	8AG	6.3	0.3	—	—	—	Indicator Tube											6AF7G
6AG6G	Power-Amplifier Pentode	7S	6.3	1.25	—	—	—	Class-A Amplifier	250	− 6.0	250	6.0	32	—	10000	—	8500	3.75	6AG6G
6AH5G	Beam Power Amplifier	6AP	6.3	0.9	—	—	—	Class-A Amplifier	350	−18	250	—	—	33000	5200	—	4200	10.8	6AH5G
6AH7GT	Twin Triode	8BE	6.3	0.3	—	—	—	Converter & Amp.	250	− 9.0	—	—	12[1]	6600	2400	16	—	—	6AH7GT
6AL6G	Beam Power Amplifier	6AM	6.3	0.9	—	—	—	Class-A Amplifier	250	−14.0	250	5.0	72	22500	6000	—	2500	6.5	6AL6G
6AL7GT	Electron-Ray Tube	8CH	6.3	0.15	—	—	—	Indicator	Outer edge of any of the three illuminated areas displaced 1/16 in. min. outward with +5 volts to its electrode. Similar inward disp. with —5 volts. No pattern with —6 volts grid.										6AL7GT
6AQ7GT	Duplex Diode Triode	8CK	6.3	0.3	2.3	1.5	2.8	Class-A Amplifier	250	− 2.0	—	—	2.3	44000	1600	70	—	—	6AQ7GT
6AR6	Beam Power Amp.	6BQ	6.3	1.2	11	7	0.55	Class-A Amplifier	250	−22.5	250	5	77	21000	5400	95	—	—	6AR6
6AR7GT	Diode Triode	8CG	6.3	0.3	1.4	1	2	Class-A Amplifier	250	− 2	—	—	1.3	66500	1050	70	—	—	6AR7GT
6AS7G	Low-Mu Twin Triode	8BD	6.3	2.5	—	—	—	D.C. Amplifier	135	250*	—	—	125	280	7500	2.1	—	—	6AS7G
								Class-A_1 Amp. P.P.	250	2500*	—	—	100/106	280	225[9]	—	6000[6]	13	
6B4G	Triode Power Amplifier	5S	6.3	1.0	—	—	—	Power Amplifier		Characteristics same as Type 6A3—Table IV						—	—	—	6B4G
6B6G	Duplex-Diode High-μ Triode	7V	6.3	0.3	1.7	3.8	1.7	Detector-Amplifier		Characteristics same as Type 75—Table IV						—	—	—	6B6G
6BQ6GT	Beam Pentode	6AM	6.3	1.2	—	—	—	Deflection Amp.	250	47*	150	2.1	45		5500	—	—	—	6BQ6GT
6BG6	Beam Power Amplifier	5BT	6.3	0.9	11	6.5	0.5	Deflection Amp.	400	−50	350	6.0	70	—	6000	—	—	—	6BG6
6C8G	Twin Triode	8G	6.3	0.3	—	—	—	Amp. 1 Section	250	− 4.5	—	—	3.1	26000	1450	38	—	—	6C8G
6D8G	Pentagrid Converter	8A	6.3	0.15	—	—	—	Converter	250	− 3.0	100	Cathode current 13.0Ma.			Anode grid (No. 2) Volts = 250[3]				6D8G
6E8G[10]	Triode-Hexode Converter	8O	6.3	0.3	—	—	—	Converter	250	− 2.0			Triode Plate 150 volts						6E8G
6F8G	Twin Triode	8G	6.3	0.6	—	—	—	Amplifier	250	− 8.0	—	—	9[1]	7700	2600	20	—	—	6F8G
6G6G	Pentode Power Amplifier	7S	6.3	0.15	—	—	—	Class-A Amplifier	180	− 9.0	180	2.5	15	175000	2300	400	10000	1.1	6G6G
								Class-A Amplifier[2]	180	−12.0	—	—	—	4750	2000	9.5	12000	0.25	
6H4GT	Diode Rectifier	5AF	6.3	0.15	—	—	—	Detector	100	—	—	—	4.0	—	—	—	—	—	6H4GT
6H8G	Duo-Diode High-μ Pentode	8E	6.3	0.3	—	—	—	Class-A Amplifier	250	− 2.0	100	—	8.5	650000	2400	—	—	—	6H8G
6J8G[10]	Triode Heptode	8H	6.3	0.3	—	—	—	Converter	250	− 3.0	100	2.8	1.2	Anode-grid (No. 2) 250 volts max.[3] 5 ma.					6J8G
6K5GT[10]	High-μ Triode	5U	6.3	0.3	2.4	3.6	2.0	Class-A Amplifier	250	− 3.0	—	—	1.1	50000	1400	70	—	—	6K5GT
6K6GT	Pentode Power Amplifier	7S	6.3	0.4	—	—	—	Class-A Amplifier	Characteristics same as Type 41—Table IV										6K6GT
6L5G	Triode Amplifier	6Q	6.3	0.15	2.8	5.0	2.8	Class-A Amplifier	250	− 9.0	—	—	8.0	—	1900	17	—	—	6L5G
6M6G	Power Amplifier Pentode	7S	6.3	1.2	—	—	—	Class-A Amplifier	250	− 6.0	250	4.0	36	—	9500	—	7000	4.4	6M6G
6M7G	Pentode Amplifier	7R	6.3	0.3	—	—	—	R.F. Amplifier	250	− 2.5	125	2.8	10.5	900000	3400	—	—	—	6M7G
6M8GT	Diode Triode Pentode	8AU	6.3	0.6	—	—	—	Triode Amplifier	100	—	—	—	0.5	91000	1100	—	—	—	6M8GT
								Pentode Amplifier	100	− 3.0	100	—	8.5	200000	1900	—	—	—	
6N6G[10]	Direct-Coupled Amplifier	7AU	6.3	0.8	—	—	—	Power Amplifier	Characteristics same as Type 6B5—Table IV							—	—	—	6N6G
6P5GT[10]	Triode Amplifier	6Q	6.3	0.3	3.4	5.5	2.6	Class-A Amplifier	250	−13.5	—	—	5.0	9500	1450	13.8	—	—	6P5GT
6P7G[10]	Triode-Pentode	7U	6.3	0.3	—	—	—	Class-A Amplifier	Characteristics same as 6F7—Table IV										6P7G
6P8G	Triode-Hexode Converter	8K	6.3	0.8	—	—	—	Converter	250	− 2.0	75	1.4	1.5	Triode Plate 100 v. 2.2 ma.					6P8G
6Q6G	Diode-Triode	6Y	6.3	0.15	—	—	—	Class-A Amplifier	250	− 3.0	—	—	1.2	—	1050	65	—	—	6Q6G
6R6G	Pentode Amplifier	6AW	6.3	0.3	4.5	11	0.007	Class-A Amplifier	250	− 3.0	100	1.7	7.0	—	1450	1160	—	—	6R6G
6S6GT	Remote Cut-off Pentode	5AK	6.3	0.45	—	—	—	R.F. Amplifier	250	− 2.0	100	3.0	13	350000	4000	—	—	—	6S6GT
6S8GT	Triple Diode Triode	8CB	6.3	0.3	1.2	5	2	Class-A Amplifier	250	− 2.0	—	—	0.9	91000	1100	100	—	—	6S8GT
6SD7GT	Medium Cut-off Pentode	8M	6.3	0.3	9	7.5	.0035	R.F. Amplifier	250	− 2.0	100	1.9	6.0	1000000	3600	—	—	—	6SD7GT
6SE7GT	Sharp Cut-off Pentode	8N	6.3	0.3	8	7.5	.005	R.F. Amplifier	250	− 1.5	100	1.5	4.5	1100000	3400	3750	—	—	6SE7GT
6SH7L	Pentode R.F. Amp.	8BK	6.3	0.3	—	—	—	Class-A Amplifier	100	− 1.0	100	2.1	5.3	350000	4000	—	—	—	6SH7L
									250	− 1.0	150	4.1	10.8	900000	4900	—	—	—	
6SL7GT	Twin Triode	8BD	6.3	0.3	—	—	—	Class-A Amplifier	250	− 2.0	—	—	2.3[1]	44000	1600	70	—	—	6SL7GT
6SN7GT	Twin Triode	8BD	6.3	0.6	—	—	—	Class-A Amplifier	250	− 8.0	—	—	9.0[1]	7700	2600	20	—	—	6SN7GT

TABLE II—6.3-VOLT GLASS TUBES WITH OCTAL BASES—*Continued*

Type	Name	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
6SU7GTY	Twin Triode	8BD	6.3	0.3	—	—	—	Class-A Amplifier	250	−2.0	—	—	2.3	44000	1600	70	—	—	6SU7GTY
6T6GM [10]	Amplifier	6Z	6.3	0.45	—	—	—	Class-A Amplifier	250	−1.0	100	2.0	10	1000000	5500	—	—	—	6T6GM
6U6GT	Beam Power Amplifier	7AC	6.3	0.75	—	—	—	Class-A Amplifier	200	−14.0	135	3.0	56	20000	6200	—	3000	5.5	6U6GT
6U7G	Variable-μ Pentode	7R	6.3	0.3	5	9	.007	Class-A Amplifier		Characteristics same as Type 6D6—Table III									6U7G
6V7G [10]	Duplex Diode-Triode	7V	6.3	0.3	2	3.5	1.7	Detector-Amplifier		Characteristics same as Type 85—Table III									6V7G
6W6GT	Beam Power Amplifier	7AC	6.3	1.25	—	—	—	Class-A Amplifier	135	−9.5	135	12.0	61.0	—	9000	215	2000	3.3	6W6GT
6W7G	Pentode Det. Amplifier	7R	6.3	0.15	5	8.5	.007	Class-A Amplifier	250	−3.0	100	2.0	0.5	1500000	1225	1850	—	—	6W7G
6X6G	Electron-Ray Tube	7AL	6.3	0.3	—	—	—	Indicator Tube	250	0 v. for 300°, 2 ma. —8 v. for 0°, 0 ma. Vane grid 125 v.									6X6G
6Y6G	Beam Power Amplifier	7AC	6.3	1.25	15	8	0.7	Class-A Amplifier	135	−13.5	135	3.0	60.0	9300	7000	—	2000	3.6	6Y6G
6Y7G [10]	Twin Triode Amplifier	8B	6.3	0.3	—	—	—	Class-B Amplifier		Characteristics same as Type 79—Table IV									6Y7G
6Z7G	Twin Triode Amplifier	8B	6.3	0.3	—	—	—	Class-B Amplifier	180	0	—	—	8.4	—	—	—	12000	4.2	6Z7G
									135	0	—	—	6.0	—	—	—	9000	2.5	
717A	Sharp Cut-off Pentode	8BK	6.3	0.175	—	—	—	Class-A Amplifier	120	−2.0	120	2.5	7.5	390000	4000	—	—	—	717A
1223	Sharp Cut-off Pentode	7R	6.3	0.3	—	—	—	Class-A Amplifier		Characteristics same as 6C6—Table IV									1223
1635	Twin Triode Amplifier	8B	6.3	0.6	—	—	—	Class-B Amplifier	400	0	—	—	10/63	—	—	—	14000	17	1635
5691	Hi-Mu Twin Triode	8BD	6.3	0.6	2.4 [7] 2.7 [8]	2.3 [7] 2.7 [8]	3.6 [7] 3.6 [8]	Class-A Amp.	250	−2	—	—	2.3 [1]	44000	1600	70	—	—	5691
5692	Medium-Mu Twin Triode	8BD	6.3	0.6	2.3 [7] 2.6 [8]	2.5 [7] 2.7 [8]	3.5 [7] 3.3 [8]	Class-A Amp.	250	−9	—	—	6.5 [1]	9100	2200	18	—	—	5692
7000	Low-Noise Amplifier	7R	6.3	0.3	—	—	—	Class-A Amplifier		Characteristics same as Type 6J7—Table									7000

* Cathode resistor-ohms.
[1] Per plate.
[2] Screen tied to plate.
[3] Through 20,000-ohm dropping resistor.
[4] Values are for single tube.
[5] Values are for two tubes in push-pull.
[6] Plate-to-plate value.
[7] No. 1 triode.
[8] No. 2 triode.
[9] Peak a.f. volts G-G.
[10] Discontinued.

TABLE III—7-VOLT LOCK-IN-BASE TUBES

For other lock-in-base types see Tables VIII, IX, and X

Type	Name	Socket Connections	Heater Volts	Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
7A4	Triode Amplifier	5AC	7.0	0.32	3.4	3	4	Class-A Amplifier	250	−8.0	—	—	9.0	7700	2600	20	—	—	7A4
7A5	Beam Power Amplifier	6AA	7.0	0.75	13	7.2	0.44	Class-A_1 Amplifier	125	−9.0	125	3.2/8	37.5/40	17000	6100	—	2700	1.9	7A5
7A6	Twin Diode	7AJ	7.0	0.16	—	—	—	Rectifier			Max. A.C. volts per plate—150. Max. Output current—10 ma.								7A6
7A7	Remote Cut-off Pentode	8V	7.0	0.32	6	7	.005	Class-A Amplifier	250	−3.0	100	2.0	8.6	800000	2000	1600	—	—	7A7
7A8	Multigrid Converter	8U	7.0	0.16	7.5	9.0	0.15	Converter	250	−3.0	100	3.1	3.0	50000	Anode-grid 250 volts max.[1]				7A8
7AD7	Pentode	8V	6.3	0.6	11.5	7.5	0.03	Class-A_1 Amp.	300	68*	150	7.0	28.0	300000	9500	—	—	—	7AD7
7AF7	Twin Triode	8AC	6.3	0.3	2.2	1.6	2.3	Class-A Amp.	250	−10	—	—	9.0	7600	2100	16	—	—	7AF7
7AG7	Sharp Cut-off Pentode	8V	7.0	0.16	7.0	6.0	0.005	Class-A_1 Amp.	250	250*	250	2.0	6.0	750000	4200	—	—	—	7AG7
7AH7	Pentode Amplifier	8V	6.3	0.15	7.0	6.5	0.005	Class-A_1 Amplifier	250	250*	250	1.9	6.8	1000000	3300	—	—	—	7AH7
7B4	High-μ Triode	5AC	7.0	0.32	3.6	3.4	1.6	Class-A Amplifier	250	−2.0	—	—	0.9	66000	1500	100	—	—	7B4
7B5	Pentode Power Amplifier	6AE	7.0	0.43	3.2	3.2	1.6	Class-A_1 Amplifier	250	−18.0	250	5.5/10	32/33	68000	2300	—	7600	3.4	7B5
7B6	Duo-Diode Triode	8W	7.0	0.32	3.0	2.4	1.6	Class-A Amplifier	250	−2.0	—	—	1.0	91000	1100	100	—	—	7B6
7B7	Remote Cut-off Pentode	8V	7.0	0.16	5	7	.005	Class-A Amplifier	250	−3.0	100	2.0	8.5	700000	1700	1200	—	—	7B7
7B8	Pentagrid Converter	8X	7.0	0.32	10.0	9.0	0.2	Converter	250	−3.0	100	2.7	3.5	360000	Anode-grid 250 volts max.[1]				7B8
7C5	Tetrode Power Amplifier	6AA	7.0	0.48	9.5	9.0	0.4	Class-A_1 Amplifier	250	−12.5	250	4.5/7	45/47	52000	4100	—	5000	4.5	7C5
7C6	Duo-Diode Triode	8W	7.0	0.16	2.4	3	1.4	Class-A Amplifier	250	−1.0	—	—	1.3	100000	1000	100	—	—	7C6
7C7	Pentode Amplifier	8V	7.0	0.16	5.5	6.5	.007	Class-A Amplifier	250	−3.0	100	0.5	2.0	2 meg.	1300	—	—	—	7C7
7D7	Triode-Hexode Converter	8AR	7.0	0.48	—	—	—	Converter	250	−3.0	Triode Plate (No. 3) 150 v. 3.5 ma.								7D7

TABLE III—7-VOLT LOCK-IN-BASE TUBES—*Continued*

Type	Name	Socket Connec-tions	Heater Volts	Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transcon-ductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
7E6	Duo-Diode Triode	8W	7.0	0.32	—	—	—	Class-A Amplifier	250	− 9.0	—	—	9.5	8500	1900	16	—	—	7E6
7E7	Duo-Diode Pentode	8AE	7.0	0.32	4.6	4.6	.005	Class-A Amplifier	250	− 3.0	100	1.6	7.5	700000	1300	—	—	—	7E7
7F7	Twin Triode	8AC	7.0	0.32	—	—	—	Class-A Amplifier [2]	250	− 2.0	—	—	2.3	44000	1600	70	—	—	7F7
7F8	Twin Triode	8BW	6.3	0.30	2.8	1.4	1.2	R.F. Amplifier	250 180	− 2.5 − 1.0	—	—	10.0 12.0	10400 8500	5000 7000	—	—	—	7F8
7G7/ 1232	Sharp Cut-off Pentode	8V	7.0	0.48	9	7	.007	Class-A Amplifier	250	− 2.0	100	2.0	6.0	800000	4500	—	—	—	7G7/ 1232
7G8/ 1206	Dual Tetrode	8BV	6.3	0.30	3.4	2.6	0.15	R.F. Amplifier [2]	250	− 2.5	100	0.8	4.5	225000	2100	—	—	—	7G8/ 1206
7H7	Semi-Variable-μ Pentode	8V	7.0	0.32	8	7	.007	R.F. Amplifier	250	− 2.5	150	2.5	9.0	1000000	3500	—	—	—	7H7
7J7	Triode-Heptode Converter	8AR	7.0	0.32	—	—	—	Converter	250	− 3.0	100	2.9	1.3	Triode Plate 250 v. Max.[1]					7J7
7K7	Duo-Diode High-μ Triode	8BF	7.0	0.32	—	—	—	Class-A Amplifier	250	− 2.0	—	—	2.3	44000	1600	70	—	—	7K7
7L7	Sharp Cut-off Pentode	8V	7.0	0.32	8	6.5	.01	Class-A Amplifier	250	− 1.5	100	1.5	4.5	100000	3100	Cathode Resistor 250 ohms			7L7
7N7	Twin Triode	8AC	7.0	0.6	3.4 [3] 2.9 [4]	2.0 [3] 2.4 [4]	3.0 [3] 3.0 [4]	Class-A Amplifier [2]	250	− 8.0	—	—	9.0	7700	2600	20	—	—	7N7
7Q7	Pentagrid Converter	8AL	7.0	0.32	—	—	—	Converter	250	0	100	8.0	3.4	800000	Grid No. 1 resistor 20000 ohms				7Q7
7R7	Duo-Diode Pentode	8AE	7.0	0.32	5.6	5.3	.004	Class-A Amplifier	250	− 1.0	100	1.7	5.7	1000000	3200	—	—	—	7R7
7S7	Triode Hexode Converter	8BL	7.0	0.32	—	—	—	Converter	250	− 2.0	100	2.2	1.7	2000000	Triode Plate 250 v. Max.[1]				7S7
7T7	Pentode Amplifier	8V	7.0	0.32	8	7	.005	Class-A Amplifier	250	− 1.0	150	4.1	10.8	900000	4900	—	—	—	7T7
7V7	Sharp Cut-off Pentode	8V	7.0	0.48	9.5	6.5	.004	Class-A Amplifier	300	160*	150	3.9	10	300000	5800	—	—	—	7V7
7W7	Sharp Cut-off Pentode	8BJ	7.0	0.48	9.5	7.0	.0025	Class-A Amplifier	300	− 2.2	150	3.9	10	300000	5800	—	—	—	7W7
7X7	Duo-Diode Triode	8BZ	6.3	0.3	—	—	—	Class-A Amplifier	250	− 1.0	—	—	1.9	67000	1500	100	—	—	7X7
1231	Pentode Amplifier	8V	6.3	0.45	8.5	6.5	.015	Class-A Amplifier	300	200*	150	2.5	10	700000	5500	3850	—	—	1231
1273	Nonmicrophonic Pentode	8V	7.0	0.32	6.0	6.5	.007	Class-A_1 Amplifier	250 100	− 3.0 − 1.0	100 100	0.7 1.8	2.2 5.7	1000000 400000	1575 2275	— —	— —	— —	1273
5679	Twin Diode	7CX	6.3	0.15	—	—	—	V.T.V.M. Rectifier	Same as 7A6										5679
XXL	Triode Oscillator	5AC	7.0	0.32	—	—	—	Oscillator	250	− 8.0	—	—	8.0	—	2300	20	—	—	XXL

* Cathode resistor—ohms. [1] Applied through 20000-ohm dropping resistor. [2] Each section. [3] Triode No. 1. [4] Triode No. 2.

TABLE IV—6.3-VOLT GLASS RECEIVING TUBES

Type	Name	Base	Socket Connec-tions	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transcon-ductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
2C21/ 1642	Twin-Triode Amplifier	M.	7BH	6.3	0.6	—	—	—	Class-A Amp.	250	−16.5	—	—	8.3	7600	1375	10.4	—	—	2C21/ 1642
6A3	Triode Power Amplifier	M.	4D	6.3	1.0	7.0	5.0	16.0	Class-A Amp.	250	−45	—	—	60	800	5250	4.2	2500	3.5	6A3
									Class AB_1 Amp.[10]	300 300	−62 850*	Fixed Bias Self Bias		80 80	—	—	—	3000 [11] 5000 [11]	15 10	
6A4#	Pentode Power Amplifier	M.	5B	6.3	0.3	—	—	—	Class-A Amp.	180	−12.0	180	3.9	22	60000	2500	150	8000	1.5	6A4
6A6	Twin Triode Amplifier	M.	7B	6.3	0.8	—	—	—	Class-B Amp. P.P.	250 300	0 0	—	—	Power output is for one tube at stated load, plate-to-plate				8000 10000	8.0 10.0	6A6
6A7	Pentagrid Converter	S.	7C	6.3	0.3	8.5	9.0	0.3	Converter	250	− 3.0	100	2.2	3.5	360000	Anode grid (No. 2) 200 volts max.				6A7
6AB5/6N5	Electron-Ray Tube	S.	6R	6.3	0.15	—	—	—	Indicator Tube	180	Cut-off Grid Bias = −12 v.			0.5	Target Current 2 ma.				—	6AB5/6N5
6AF6G	Electron-Ray Tube Twin Indicator Type	S.	7AG	6.3	0.15	—	—	—	Indicator Tube	135 100	Ray Control Voltage = 81 for 0° Shadow Angle. Target current 1.5 ma. Ray Control Voltage = 60 for 0° Shadow Angle. Target current 0.9 ma.									6AF6G
6B5	Direct-Coupled Power Amplifier	M.	6AS	6.3	0.8	—	—	—	Class-A Amp.[9] Push-Pull Amp.[10]	300 400	0 −13.0	— —	6 [1] 4.5 [1]	45 40	241000 —	2400 —	58 —	7000 10000 [11]	4.0 20	6B5

TABLE IV—6.3-VOLT GLASS RECEIVING TUBES—*Continued*

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
6B7	Duplex-Diode Pentode	S.	7D	6.3	0.3	3.5	9.5	.007	Pentode R.F. Amp.	250	−3.0	125	2.3	9.0	650000	1125	730	—	—	6B7
6C6	Sharp Cut-off Pentode	S.	6F	6.3	0.3	5	6.5	.007	R.F. Amplifier	250	−3.0	100	0.5	2.0	1500000	1225	1500	—	—	6C6
6C7 #	Duplex Diode Triode	S.	7G	6.3	0.3	—	—	—	Class-A Amp.	250	−9.0	—	—	4.5	—	20	1250	—	—	6C7
6D6	Variable-μ Pentode	S.	6F	6.3	0.3	4.7	6.5	.007	R.F. Amplifier	250	−3.0	100	2.0	8.2	800000	1600	1280	—	—	6D6
6D7 #	Sharp Cut-off Pentode	S.	7H	6.3	0.3	5.2	6.8	.01	Class-A Amp.	250	−3.0	100	0.5	2.0	—	1600	1280	—	—	6D7
6E5	Electron-Ray Tube	S.	6R	6.3	0.3	—	—	—	Indicator Tube	250	0	—	—	0.25	Target Current 4 ma.				—	6E5
6E6 #	Twin Triode Amplifier	M.	7B	6.3	0.6	—	—	—	Class-A Amp.	250	−27.5	Per plate—18.0			3500	1700	6.0	14000	1.6	6E6
6E7 #	Variable-μ Pentode	S.	7H	6.3	0.3	—	—	—	R.F. Amplifier				Characteristics same as 6U7G—Table II							6E7
6F7	Triode Pentode	S.	7E	6.3	0.3	—	—	—	Triode Unit Amp.	100	−3.0	—	—	3.5	16000	500	8	—	—	6F7
									Pentode Unit Amplifier	250	−3.0	100	1.5	6.5	850000	1100	900	—	—	
6U5/6G5	Electron-Ray Tube	S.	6R	6.3	0.3	—	—	—	Indicator Tube	250 100	Cut-off Grid Bias = −22 v. Cut-off Grid Bias = −8 v.			0.24 0.19	Target Current 4 ma. Target Current 1 ma.				—	6U5/6G5
6H5	Electron-Ray Tube	S.	6R	6.3	0.3	—	—	—	Indicator Tube			Same characteristics as Type 6G5—Circular Pattern								6H5
6T5	Electron-Ray Tube	S.	6R	6.3	0.3	—	—	—	Indicator Tube	250	Cut-off Grid Bias = −12 v.			0.24	Target Current 4 ma.				—	6T5
36	Tetrode R.F. Amplifier	S.	5E	6.3	0.3	3.8	9	.007	R.F. Amplifier	250	−3.0	90	1.7	3.2	550000	1080	595	—	—	36
37	Triode Detector Amplifier	S.	5A	6.3	0.3	3.5	2.9	2	Class-A Amp.	250	−18.0	—	—	7.5	8400	1100	9.2	—	—	37
38	Pentode Power Amplifier	S.	5F	6.3	0.3	3.5	7.5	0.3	Class-A Amp.	250	−25.0	250	3.8	22.0	100000	1200	120	10000	2.5	38
39/44	Remote Cut-off Pentode	S.	5F	6.3	0.3	3.8	10	.007	R.F. Amplifier	250	−3.0	90	1.4	5.8	1000000	1050	1050	—	—	39/44
41	Pentode Power Amplifier	S.	6B	6.3	0.4	—	—	—	Class-A Amp.	250	−18.0	250	5.5	32.0	68000	2200	150	7600	3.4	41
42	Pentode Power Amplifier	M.	6B	6.3	0.7	—	—	—	Class-A Amp.	250	−16.5	250	6.5	34.0	100000	2200	220	7000	3.0	42
52	Dual Grid Triode	M.	5C	6.3	0.3	—	—	—	Class-A Amp.[4]	110	0	—	—	43.0	1750	3000	5.2	2000	1.5	52
									Class-B, 2 tubes[5]	180	0	—	—	3.0[12]				10000	5.0	
56AS	Triode Amplifier	S.	5A	6.3	0.4	—	—	—	Class-A Amp.	Characteristics same as 56										56AS
57AS	Sharp Cut-off Pentode	S.	6F	6.3	0.4	—	—	—	R.F. Amplifier	Characteristics same as 57										57AS
58AS	Remote Cut-off Pentode	S.	6F	6.3	0.4	—	—	—	R.F. Amplifier	Characteristics same as 58										58AS
75	Duplex-Diode Triode	S.	6G	6.3	0.3	1.7	3.8	1.	Triode Amplifier	250	−1.35	—	—	0.4	91000	1100	100	—	—	75
76	Triode Detector Amplifier	S.	5A	6.3	0.3	3.5	2.5	2.8	Class-A Amp.	250	−13.5	—	—	5.0	9500	1450	13.8	—	—	76
77	Sharp Cut-off Pentode	S.	6F	6.3	0.3	4.7	11	.007	R.F. Amplifier	250	−3.0	100	0.5	2.3	1500000	1250	1500	—	—	77
78	Variable-μ Pentode	S.	6F	6.3	0.3	4.5	11	.007	R.F. Amplifier	250	−3.0	100	1.7	7.0	800000	1450	1160	—	—	78
79	Twin Triode Amplifier	S.	6H	6.3	0.6	—	—	—	Class-B Amp.	250	0	—	—	10.6[12]	Power output is for one tube			14000	8.0	79
85	Duplex-Diode Triode	S.	6G	6.3	0.3	1.5	4.3	1.5	Class-A Amp.	250	−20.0	—	—	8.0	7500	1100	8.3	20000	0.35	85
85AS	Duplex-Diode Triode	S.	6G	6.3	0.3	—	—	—	Class-A Amp.	250	−9.0	—	—	5.5	—	1250	20	—	—	85AS
89	Power Amplifier Pentode	S.	6F	6.3	0.4	—	—	—	Triode Amp.[2]	250	−31.0	—	—	32.0	2600	1800	4.7	5500	0.9	89
									Pentode Amp.[8]	250	−25.0	250	5.5	32.0	70000	1800	125	6750	3.4	
1221	Pentode R.F. Amplifier	S.	6F	6.3	0.3	—	—	—	Class-A Amp.	Special non-microphonic. Characteristics same as 6C6										1221
1603[3]	Sharp Cut-off Pentode	M.	6F	6.3	0.3	—	—	—	Class-A Amp.	Characteristics same as 6C6										1603
7700[3]	Sharp Cut-off Pentode	S.	6F	6.3	0.3	—	—	—	Class-A Amp.	Characteristics same as 6C6										7700

* Cathode bias resistor—ohms.
\# Discontinued.

[1] Current to input plate (P_1).
[2] Grids Nos. 2 and 3 connected to plate.
[3] Low noise, nonmicrophonic tubes.
[4] G_2 tied to plate.
[5] G_1 tied to G_2.
[6] Osc. grid leak ohms.
[7] Screen dropping resistor ohms.
[8] Grid No. 2, screen; grid No. 3, suppressor.
[9] Values for single tube.
[10] Values for two tubes in push-pull.
[11] Plate-to-plate value.
[12] No signal value.

TABLE V—2.5-VOLT RECEIVING TUBES

Type	Name	Base	Socket Connec-tions	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance $\mu\mu$fd. In	Capacitance $\mu\mu$fd. Out	Capacitance $\mu\mu$fd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transcon-ductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
2S/4S	Duodiode	M.	5D	2.5	1.35	—	—	—	Detector	At 50 d.c. Volts per plate, cathode ma.=80										2S/4S
2A3	Triode Power Amplifier	M.	4D	2.5	2.5	7.5	5.5	16.5	Class-A Amp.	Characteristics same as Type 6A3, Table IV										2A3
2A5	Pentode Power Amplifier	M.	6B	2.5	1.75	—	—	—	Class-A Amp.	Characteristics same as Type 42, Table IV										2A5
2A6	Duplex-Diode Triode	S.	6G	2.5	0.8	1.7	3.8	1.7	Class-A Amp.	Characteristics same as Type 75, Table IV										2A6
2A7	Pentagrid Converter	S.	7C	2.5	0.8	—	—	—	Converter	Characteristics same as Type 6A7, Table IV										2A7
2B6	Direct-Coupled Amplifier	M.	7J	2.5	2.25	—	—	—	Amplifier	250	−24.0	—	—	40.0	5150	3500	18.0	5000	4.0	2B6
2B7	Duplex-Diode Pentode	S.	7D	2.5	0.8	3.5	9.5	.007	Pentode Amp.	Characteristics same as Type 6B7—Table IV										2B7
2E5	Electron-Ray Tube	S.	6R	2.5	0.8	—	—	—	Indicator Tube	Characteristics same as Type 6E5—Table IV										2E5
2G5	Electron-Ray Tube	S.	6R	2.5	0.8	—	—	—	Indicator Tube	Characteristics same as 6U5/6G5—Table IV										2G5
24-A	Tetrode R.F. Amplifier	M.	5E	2.5	1.75	5.3	10.5	.007	Screen-Grid R.F. Amplifier	250	− 3.0	90	1.7	4.0	600000	1050	630	—	—	24-A
									Bias Detector	250	− 5.0	20/45	Plate current adjusted to 0.1 ma. with no signal							
27	Triode Detector-Amplifier	M.	5A	2.5	1.75	3.1	2.3	3.3	Class-A Amp.	250	−21.0	—	—	5.2	9250	975	9.0	—	—	27
									Bias Detector	250	−30.0	—	Plate current adjusted to 0.2 ma. with no signal							
35/51	Remote Cut-off Pentode	M.	5E	2.5	1.75	5.3	10.5	.007	Screen-Grid R.F. Amplifier	250	− 3.0	90	2.5	6.5	400000	1050	420	—	—	35/51
45	Triode Power Amplifier	M.	4D	2.5	1.5	4	3	7	Class-A Amp.	275	−56.0	—	—	36.0	1700	2050	3.5	4600	2.00	45
46	Dual-Grid Power Amp.	M.	5C	2.5	1.75	—	—	—	Class-A Amp.[2]	250	−33.0	—	—	22.0	2380	2350	5.6	6400	1.25	46
									Class-B Amp.[3]	400	0	—	—	12	Power output for 2 tubes			5800	20.0	
47	Pentode Power Amplifier	M.	5B	2.5	1.75	8.6	13	1.2	Class-A Amp.	250	−16.5	250	6.0	31.0	60000	2500	150	7000	2.7	47
53	Twin Triode Amplifier	M.	7B	2.5	2.0	—	—	—	Class-B Amp.	Characteristics same as Type 6A6, Table IV										53
55	Duplex-Diode Triode	S.	6G	2.5	1.0	1.5	4.3	1.5	Class-A Amp.	Characteristics same as Type 85, Table IV										55
56	Triode Amplifier, Detector	S.	5A	2.5	1.0	3.2	2.4	3.2	Class-A Amp.	Characteristics same as Type 76, Table IV										56
57	Sharp Cut-off Pentode	S.	6F	2.5	1.0	—	—	—	R.F. Amplifier	250	− 3.0	100	0.5	2.0	1500000	1225	1500	—	—	57
58	Remote Cut-off Pentode	S.	6F	2.5	1.0	4.7	6.3	.007	Screen-Grid R.F. Amplifier	250	− 3.0	100	2.0	8.2	800000	1600	1280	—	—	58
59	Pentode Power Amplifier	M.	7A	2.5	2.0	—	—	—	Class-A Triode[4]	250	−28.0	—	—	26.0	2300	2600	6.0	5000	1.25	59
									Class-A Pentode[5]	250	−18.0	250	9.0	35.0	40000	2500	100	6000	3.0	
RK15	Triode Power Amplifier	M.	4D[1]	2.5	1.75	—	—	—		Characteristics same as Type 46 with Class-B connections										RK15
RK16	Triode Power Amplifier	M.	5A	2.5	2.0	—	—	—		Characteristics same as Type 59 with Class-A triode connections										RK16
RK17	Pentode Power Amplifier	M.	5F	2.5	2.0	—	—	—		Characteristics same as Type 2A5										RK17

[1] Grid connection to cap; no connection to No. 3 pin. [2] Grid No. 2 tied to plate. [3] Grids Nos. 1 and 2 tied together. [4] Grids Nos. 2 and 3 connected to plate. [5] Grid No. 2, screen; grid No. 3, suppressor.

TABLE VI—2.0-VOLT BATTERY RECEIVING TUBES

Type	Name	Base	Socket Connec-tions	Filament Volts	Filament Amp.	Capacitance $\mu\mu$fd. In	Capacitance $\mu\mu$fd. Out	Capacitance $\mu\mu$fd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transcon-ductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
1A4P	Variable-μ Pentode	S.	4M	2.0	0.06	5	11	.007	R.F. Amplifier	180	− 3.0	67.5	0.8	2.3	1000000	750	750	—	—	1A4P
1A4T	Variable-μ Tetrode	S.	4K	2.0	0.06	5	11	.007	R.F. Amplifier	180	− 3.0	67.5	0.7	2.3	960000	750	720	—	—	1A4T
1A6	Pentagrid Converter	S.	6L	2.0	0.06	—	—	—	Converter	180	− 3.0	67.5	2.4	1.3	500000	Anode grid (No. 2) 180 max. volts				1A6
1B4P/951	Pentode R.F. Amplifier	S.	4M	2.0	0.06	5	11	.007	R.F. Amplifier	180 90	− 3.0 − 3.0	67.5 67.5	0.6 0.7	1.7 1.6	1500000 1000000	650 600	1000 550	—	—	1B4P/951
1B5/25S	Duplex-Diode Triode	S.	6M	2.0	0.06	1.6	1.9	3.6	Triode Class-A	135	− 3.0	—	—	0.8	35000	575	20	—	—	1B5/25S

TABLE VI—2.0-VOLT BATTERY RECEIVING TUBES—*Continued*

Type	Name	Base	Socket Connections	Filament Volts	Filament Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
1C6	Pentagrid Converter	S.	6L	2.0	0.12	10	10	—	Converter	180	−3.0	67.5	2.0	1.5	750000	Anode grid (No. 2) 135 max. volts				1C6
1F4	Pentode Power Amplifier	M.	5K	2.0	0.12	—	—	—	Class-A Amp.	135	−4.5	135	2.6	8.0	200000	1700	340	16000	0.34	1F4
1F6	Duplex-Diode Pentode	S.	6W	2.0	0.06	4	9	.007	R.F. Amplifier	180	−1.5	67.5	0.6	2.0	1000000	650	650	—	—	1F6
									A.F. Amplifier	135	−1.0	135	Plate, 0.25 megohm; screen, 1.0 megohm					Amp. = 48		
15 #	Sharp Cut-off Pentode	S.	5F	2.0	0.22	2.3	7.8	0.01	R.F. Amplifier	135	−1.5	67.5	0.3	1.85	800000	750	600	—	—	15
19	Twin-Triode Amplifier	S.	6C	2.0	0.26	—	—	—	Class-B Amp.	135	0	—	—	—	Load plate-to-plate			10000	2.1	19
30	Triode Detector Amplifier	S.	4D	2.0	0.06	—	—	—	Class-A Amp.	180	−13.5	—	—	3.1	10300	900	9.3	—	—	30
31	Triode Power Amplifier	S.	4D	2.0	0.13	3.5	2.7	5.7	Class-A Amp.	180	−30.0	—	—	12.3	3600	1050	3.8	5700	0.375	31
32	Sharp Cut-off Pentode	M.	4K	2.0	0.06	5.3	10.5	.015	R.F. Amplifier	180	−3.0	67.5	0.4	1.7	1200000	650	780	—	—	32
33	Pentode Power Amplifier	M.	5K	2.0	0.26	8	12	1	Class-A Amp.	180	−18.0	180	5.0	22.0	55000	1700	90	6000	1.4	33
34	Variable-μ Pentode	M.	4M	2.0	0.06	6	11	.015	R.F. Amplifier	180	−3.0	67.5	1.0	2.8	1000000	620	620	—	—	34
49	Dual-Grid Power Amp.	M.	5C	2.0	0.12	—	—	—	Class-A Amp.[1]	135	−20.0	—	—	6.0	4175	1125	4.7	11000	0.17	49
									Class-B Amp.[2]	180	0	—	—	Power output for 2 tubes				12000	3.5	
840	Pentode	S.	5J	2.0	0.13	—	—	—	Class-A Amp.	180	−3.0	67.5	0.7	1.0	1000000	400	400	—	—	840
950	Pentode Power Amplifier	M.	5K	2.0	0.12	—	—	—	Class-A Amp.	135	−16.5	135	2.0	7.0	100000	1000	125	13500	0.575	950
RK24	Triode	M.	4D	2.0	0.12	—	—	—	Class-A Amp.	180	−13.5	—	—	8.0	5000	1600	8.0	12000	0.25	RK24
1229	Tetrode	M.	4K	2.0	0.06	—	—	—		Special Type 32 for low grid-current applications										1229
1230	Triode	M.	4D	2.0	0.06	3.0	2.1	6.0		Special Type 30 for low grid-current applications										1230

Discontinued. [1] Grid No. 2 tied to plate. [2] Grids Nos. 1 and 2 tied together.

TABLE VII—2.0-VOLT BATTERY TUBES WITH OCTAL BASES

Type	Name	Socket Connections	Filament Volts	Filament Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
1C7G	Heptode	7Z	2.0	0.06	10	14	0.26	Converter	Characteristics same as Type 1C6—Table VI										1C7G
1D5GP	Variable-μ Pentode	5Y	2.0	0.06	5	11	.007	R.F. Amplifier	Characteristics same as Type 1A4P—Table VI										1D5GP
1D5GT #	Variable-μ Tetrode	5R	2.0	0.06	—	—	—	R.F. Amplifier	180	−3.0	67.5	0.7	2.2	600000	650				1D5GT
1D7G	Pentagrid Converter	7Z	2.0	0.06	10.5	9.0	0.25	Converter	Characteristics same as Type 1A6—Table VI										1D7G
1E5GP	Pentode Amplifier	5Y	2.0	0.06	5	11	.007	R.F. Amplifier	Characteristics same as Type 1B4—Table VI										1E5GP
1E7G	Double Pentode Power Amp.	8C	2.0	0.24	—	—	—	Class-A Amplifier	135	−7.5	135	2.0[1]	6.5[1]	220000	1600	350	24000	0.65	1E7G
1F5G	Pentode Power Amplifier	6X	2.0	0.12	—	—	—	Class-A Amplifier	Characteristics same as Type 1F4—Table VI										1F5G
1F7G[2]	Duplex-Diode Pentode	7AD	2.0	0.06	3.8	9.5	0.01	Detector-Amplifier	Characteristics same as Type 1F6—Table VI										1F7G
1G5G	Pentode Power Amplifier	6X	2.0	0.12	—	—	—	Class-A Amplifier	135	−13.5	135	2.5	8.7	160000	1550	250	9000	0.55	1G5G
1H4G	Triode Amplifier	5S	2.0	0.06	—	—	—	Detector-Amplifier	Characteristics same as Type 30—Table VI										1H4G
1H6G	Duplex-Diode Triode	7AA	2.0	0.06	1.6	1.9	3.6	Detector-Amplifier	Characteristics same as Type 1B5—Table VI										1H6G
1J5G #	Pentode Power Amplifier	6X	2.0	0.12	—	—	—	Class-A Amplifier	135	−16.5	135	2.0	7.0	—	950	100	13500	0.45	1J5G
1J6G	Twin Triode	7AB	2.0	0.24	—	—	—	Class-B Amplifier	Characteristics same as Type 19—Table VI										1J6G
4A6G	Twin Triode	8L	2.0	0.12	—	—	—	Class-A, 1 section	90	−1.5	—	—	1.1	26600	750	20			4A6G
			4.0	0.06				Class-B, 2 sections	90	−1.5	—	—	10.8[3]				8000	1.0	

Discontinued. [1] Total current for both sections; no signal. [2] Type GV has 7AF base. [3] Max. signal.

TABLE VIII—1.5-VOLT FILAMENT BATTERY TUBES

See also Table X for Special 1.4-volt Tubes

Type	Name	Base	Socket Connections	Filament Volts	Filament Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output M-watts	Type
1A5GT	Pentode Power Amplifier	O.	6X	1.4	0.05	—	—	—	Class-A_1 Amp.	90	−4.5	90	0.8	4.0	300000	850	240	25000	115	1A5GT
1A7GT	Pentagrid Converter	O.	7Z	1.4	0.05	—	—	—	Converter	90	0	45	0.6	0.55	600000	Anode-grid volts 90				1A7GT
1AB5	Pentode R.F. Amplifier	L.	5BF	1.2	0.05	2.8	4.2	0.25	R.F. Amplifier	90 150	0 −1.5	90 150	0.8 2.0	3.5 6.8	275000 125000	1100 1350	—	—	—	1AB5
1B7GT #	Heptode	O.	7Z	1.4	0.1	—	—	—	Converter	90	0	45	1.3	1.5	350000	Grid No. 1 resistor 200,000 ohms				1B7GT
1B8GT	Diode Triode Pentode	O.	8AW	1.4	0.1	—	—	—	Triode Amplifier Pentode Amp.	90 90	0 −6.0	— 90	— 1.4	0.15 6.3	240000 —	275 1150	— —	— 14000	— 210	1B8GT
1C5GT	Pentode Power Amplifier	O.	6X	1.4	0.1	—	—	—	Class-A_1 Amp.	90	−7.5	90	1.6	7.5	115000	1550	165	8000	240	1C5GT
1D8GT	Diode Triode Pentode	O.	8AJ	1.4	0.1	—	—	—	Triode Amp. Pentode Amp.	90 90	0 −9.0	— 90	— 1.0	1.1 5.0	43500 200000	575 925	25 —	— —	— —	1D8GT
1E4G	Triode Amplifier	O.	5S	1.4	0.05	2.4	6	2.40	Class-A Amp.	90 90	0 −3.0	—	—	4.5 1.5	11000 17000	1325 825	14.5 14	—	—	1E4G
1G4GT	Triode Amplifier	O.	5S	1.4	0.05	2.2	3.4	2.80	Class-A Amp.	90	−6.0	—	—	2.3	10700	825	8.8	—	—	1G4GT
1G6GT	Twin Triode	O.	7AB	1.4	0.1	—	—	—	Class-A Amp. Class-B Amp.	90 90	0 0	— —	— —	1.0 1/7	45000 34 volts input per grid	675	30	— 12000	— 675	1G6GT
1H5GT	Diode High-μ Triode	O.	5Z	1.4	0.05	1.1	6	1.00	Class-A Amp.	90	0	—	—	0.14	240000	275	65	—	—	1H5GT
1LA4	Pentode Power Amplifier	L.	5AD	1.4	0.05	—	—	—	Class-A Amp.	90	Characteristics same as 1A5GT									1LA4
1LA6	Pentagrid Converter	L.	7AK	1.4	0.05	—	—	—	Converter	90	0	45	0.6	0.55	Anode Grid Volts 90					1LA6
1LB4	Pentode Power Amplifier	L.	5AD	1.4	0.05	—	—	—	Class-A Amp.	90	−9	90	1.0	5.0	200000	925	—	12000	200	1LB4
1LB6	Heptode Converter	L.	8AX	1.4	0.05	—	—	—	Converter	90	0	67.5	2.2	0.4	Grid No. 4—67.5 v., No. 5—0 v.					1LB6
1LC5	Remote Cut-off Pentode	L.	7AO	1.4	0.05	3.2	7	.007	R.F. Amplifier	90	0	45	0.2	1.15	1500000	775	—	—	—	1LC5
1LC6	Pentagrid Converter	L.	7AK	1.4	0.05	—	—	—	Converter	90	0	35[1]	0.7	0.75		Anode Grid Volts 45				1LC6
1LD5	Diode Pentode	L.	6AX	1.4	0.05	3.2	6	0.18	Class-A Amp.	90	0	45	0.1	0.6	950000	600	—	—	—	1LD5
1LE3	Triode Amplifier	L.	4AA	1.4	0.05	1.7	3	1.70	Class-A Amp.	90 90	0 −3	—	—	4.5 1.3	11200 19000	1300 760	14.5	—	—	1LE3
1LG5	Pentode R.F. Amp.	L.	7AO	1.4	0.05	—	—	—	Class-A Amp.	90	0	45	0.4	1.7	1000000	800	—	—	—	1LG5
1LH4	Diode High-μ Triode	L.	5AG	1.4	0.05	1.1	6	1.00	Class-A Amp.	90	0	—	—	0.15	240000	275	65	—	—	1LH4
1LN5	Remote Cut-off Pentode	L.	7AO	1.4	0.05	3.4	8	.007	Class-A Amp.	90	0	90	0.3	1.2	1500000	750	—	—	—	1LN5
1N5GT	Remote Cut-off Pentode	O.	5Y	1.4	0.05	3	10	.007	Class-A Amp.	90	0	90	0.3	1.2	1500000	750	1160	—	—	1N5GT
1N6G[4]	Diode-Power-Pentode	O.	7AM	1.4	0.05	—	—	—	Class-A Amp.	90	−4.5	90	0.6	3.1	300000	800	—	25000	100	1N6G
1P5GT	Pentode	O.	5Y	1.4	0.05	3	10	.007	R.F. Amplifier	90	0	90	0.7	2.3	800000	800	640	—	—	1P5GT
1Q5GT	Tetrode Power Amplifier	O.	6AF	1.4	0.1	—	—	—	Class-A Amp.	85 90	−5.0 −4.5	85 90	1.2 1.6	7.2 9.5	70000 75000	1950 2100	—	9000 8000	250 270	1Q5GT
1R4/1294	U.h.f. Diode	L.	4AH	1.4	0.15	—	—	—	Rectifier	Max. r.m.s. voltage per plate—30					Max. d.c. output current—340 μa.					1R4/1294
1SA6GT	Medium Cut-off Pentode	O.	6CA	1.4	0.05	5.2	8.6	0.01	R.F. Amplifier	90	0	67.5	0.68	2.45	800000	970	—	—	—	1SA6GT
1SB6GT	Diode Pentode	O.	6CB	1.4	0.05	3.2	3	0.25	Class-A Amp. R.C. Amplifier	90 90	0 0	67.5 90	0.38 Screen resistor 5 meg., grid 10 meg.	1.45	700000	665	—	— 1 meg.	— 110[2]	1SB6GT
1T5GT	Beam Power Amplifier	O.	6AF	1.4	0.05	4.8	8	0.50	Class-A Amp.	90	−6.0	90	1.4	6.5	—	1150	—	14000	170	1T5GT
3B7/1291	U.h.f. Twin Triode	L.	7BE	2.8[3]	0.11	1.4	2.6	2.6	Class-A Amp.	90	0	—	—	5.2	11350	1850	21	—	—	3B7/1291
1293	U.h.f. Triode	L.	4AA	1.4	0.11	1.7	3.0	1.7	Class-A Amp.	90	0	—	—	4.7	10750	1300	14	—	—	1293
3D6/1299	U.h.f. Tetrode	L.	6BB	2.8[3]	0.11	7.5	6.5	0.30	Class-A Amp.	135	−6	90	0.7	5.7	—	2200	—	13000	500	3D6/1299
3E6	R.F. Pentode	L.	7CJ	1.4 2.8	0.10 0.05	5.5	7.5	0.007	Class-A Amp.	90	0	90	1.3	3.8	300000	2100	—	—	—	3E6
RK42	Triode Amplifier	S.	4D	1.5	0.6	—	—	—	Class-A Amp.		Characteristics same as Type 30—Table VI									RK42
RK43	Twin Triode Amplifier	S.	6C	1.5	0.12	—	—	—	Class-A Amp.	135	−3	—	—	4.5	14500	900	13	—	—	RK43

Discontinued.

[1] Through series resistor. Screen voltage must be at least 10 volts lower than oscillator anode.

[2] Voltage gain.

[3] Center-tap filament permits 1.4-volt operation.

TABLE IX—HIGH-VOLTAGE HEATER TUBES

Type	Name	Base	Socket Connections	Heater Volts	Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
12A5[8]	Pentode Power Amplifier	M.	7F	12.6 6.3	0.3 0.6	9.0	9.0	0.3	Class-A$_1$ Amp.[6]	100 180	−15 −25	100 180	3/6.5 8/14	17/19 45/48	50000 35000	1700 2400	— —	4500 3300	0.8 3.4	12A5
12A6	Beam Power Amplifier	O.	7AC	12.6	0.15	—	—	—	Class-A Amp.	250	−12.5	250	3.5	30	70000	3000	—	7500	3.4	12A6
12A7	Rectifier-Amplifier	M.	7K	12.6	0.3	—	—	—	Class-A Amp.	135	−13.5	135	2.5	9.0	102000	975	100	13500	0.55	12A7
12A8GT	Heptode	O.	8A	12.6	0.15	9.5	12	0.26	Converter	Characteristics same as 6A8—Table I										12A8GT
12AH7GT	Twin Triode	O.	8BE	12.6	0.15	Each Triode Sect.			Class-A Amp.	180	− 6.5	—	—	7.6	8400	1900	16	—	—	12AH7GT
12B6M	Diode Triode	O.	6Y	12.6	0.15	—	—	—	Class-A Amp.	250	− 2.0	—	—	0.9	91000	1100	100	—	—	12B6M
12B7ML	Pentode Amplifier	O.	8V	12.6	0.15	—	—	—	Class-A Amp.	250	− 3.0	100	2.6	9.2	800000	2000	—	—	—	12B7ML
12B8GT[8]	Triode-Pentode	O.	8T	12.6	0.3	Triode Section Pentode Section			Class-A Amp. Class-A Amp.	100 100	− 1 − 3	— 100	— 2	0.6 8	73000 170000	1500 2100	110 360	— —	— —	12B8GT
12C8	Duplex-Diode Pentode	O.	8E	12.6	0.15	6	9	.005	Class-A Amp.	Characteristics same as 6B8—Table I										12C8
12E5GT	Triode Amplifier	O.	6Q	12.6	0.15	3.4	5.5	2.60	Class-A Amp.	250	−13.5	—	—	50	—	1450	13.8	—	—	12E5GT
12F5GT	Triode Amplifier	O.	5M	12.6	0.15	1.9	3.4	2.40	Class-A Amp.	Characteristics same as 6SF5—Table I										12F5GT
12G7G	Duplex-Diode Triode	O.	7V	12.6	0.15	—	—	—	Class-A Amp.	250	− 3.0	—	—	—	58000	1200	70	—	—	12G7G
12H6	Twin Diode	O.	7Q	12.6	0.15	—	—	—	Rectifier	Characteristics same as 6H6—Table I										12H6
12J5GT	Triode Amplifier	O.	6Q	12.6	0.15	3.4	3.6	3.40	Class-A Amp.	Characteristics same as 6J5—Table I										12J5GT
12J7GT	Sharp Cut-off Pentode	O.	7R	12.6	0.15	4.2	5.0	3.8	Class-A Amp.	Characteristics same as 6J7—Table I										12J7GT
12K7GT	Remote Cut-off Pentode	O.	7R	12.6	0.15	4.6	12	.005	R.F. Amplifier	Characteristics same as 6K7—Table I										12K7GT
12K8	Triode Hexode Converter	O.	8K	12.6	0.15	—	—	—	Converter	Characteristics same as 6K8—Table I										12K8
12L8GT	Twin Pentode	O.	8BU	12.6	0.15	5	6	0.70	Class-A$_1$ Amp.	180	− 9.0	180	2.8	13.0	160000	2150	—	10000	1.0	12L8GT
12Q7GT	Duplex-Diode Triode	O.	7V	12.6	0.15	2.2	5	1.60	Class-A Amp.	Characteristics same as 6Q7—Table I										12Q7GT
12S8GT	Triple-Diode Triode	O.	8CB	12.6	0.15	2.0	3.8	1.2	Class-A Amp.	250	− 2.0	—	—	0.9	91000	1100	100	—	—	12S8GT
12SA7	Heptode	O.	8R	12.6	0.15	9.5	12	0.13	Converter	Characteristics same as 6SA7—Table I										12SA7
12SC7	Twin Triode	O.	8S	12.6	0.15	2.2	3.0	2.0	Class-A Amp.	Characteristics same as 6SC7—Table I										12SC7
12SF5	High-μ Triode	O.	6AB	12.6	0.15	4	3.6	2.40	Class-A Amp.	Characteristics same as 6SF5—Table I										12SF5
12SF7	Diode Variable-μ Pentode	O.	7AZ	12.6	0.15	5.5	6.0	.004	Class-A Amp.	Characteristics same as 6SF7—Table I										12SF7
12SG7	Medium Cut-off Pentode	O.	8BK	12.6	0.15	8.5	7.0	.003	Class-A Amp.	Characteristics same as 6SG7—Table I										12SG7
12SH7	Sharp Cut-off Pentode	O.	8BK	12.6	0.15	8.5	7.0	.003	H-F Amplifier	Characteristics same as 6SH7—Table I										12SH7
12SJ7	Sharp Cut-off Pentode	O.	8N	12.6	0.15	—	—	—	Class-A Amp.	Characteristics same as 6SJ7—Table I										12SJ7
12SK7	Remote Cut-off Pentode	O.	8N	12.6	0.15	6.0	7.0	.003	R.F. Amplifier	Characteristics same as 6SK7—Table I										12SK7
12SL7GT	Twin Triode	O.	8BD	12.6	0.15	—	—	—	Class-A Amp.	Characteristics same as 6SL7GT—Table II										12SL7GT
12SN7GT	Twin Triode	O.	8BD	12.6	0.3	—	—	—	Class-A Amp.	Characteristics same as 6SN7GT—Table II										12SN7GT
12SQ7	Duplex-Diode Triode	O.	8Q	12.6	0.15	3.2	3.0	1.60	Class-A Amp.	Characteristics same as 6SQ7—Table I										12SQ7
12SR7	Duplex-Diode Triode	O.	8Q	12.6	0.15	3.6	2.8	2.40	Class-A Amp.	Characteristics same as 6R7—Table I										12SR7
12SW7	Duplex-Diode Triode	O.	8Q	12.6	0.15	3.0	2.8	2.4	Class-A$_1$ Amp.	250	− 9	—	—	9.5	8500	1900	16	—	—	12SW7
12SX7	Twin Triode	O.	8BD	12.6	0.3	3.0	0.8	3.6	Class-A$_1$ Amp.[5]	250	− 8	—	—	9	7700	2600	20	—	—	12SX7
12SY7	Heptode Converter	O.	8R	12.6	0.15	Osc.-Grid leak 20000 ohms			Converter	250	− 2	100	8.5	3.5	1000000	450	—	—	—	12SY7
14A4	Triode Amplifier	L.	5AC	14	0.16	3.4	3.0	4.00	Class-A Amp.	Characteristics same as 7A4—Table III										14A4
14A5	Beam Power Amplifier	L.	6AA	14	0.16	—	—	—	Class-A$_1$ Amp.	250	−12.5	250	3.5/5.5	30/32	70000	3000	—	7500	2.8	14A5
14A7/ 12B7	Remote Cut-off Pentode	L.	8V	14	0.16	6.0	7.0	.005	Class-A Amp.	250	− 3.0	100	2.6	9.2	800000	2000	—	—	—	14A7/ 12B7
14AF7	Twin Triode	L.	8AC	14	0.16	2.2	1.6	2.30	Class-A Amp.	250	−10	—	—	9	7600	2100	16	—	—	14AF7
14B6	Duplex-Diode Triode	L.	8W	14	0.16	—	—	—	Class-A Amp.	Characteristics same as 7B6—Table III										14B6
14B8	Pentagrid Converter	L.	8X	14	0.16	Ic2=4 Ma.			Converter	Characteristics same as 7B8—Table III										14B8
14C5	Beam Power Amplifier	L.	6AA	14	0.24	—	—	—	Class-A Amp.	Characteristics same as 6V6—Table I										14C5

Type	Name	Base	Socket Connections	Heater Volts	Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
14C7	R.F. Pentode	L.	8V	14	0.16	6.0	6.5	.007	Class-A Amp.	250	– 3.0	100	0.7	2.2	1000000	1575	—	—	—	14C7
14E6	Duplex-Diode Triode	L.	8W	14	0.16	—	—	—	Class-A Amp.	Characteristics same as 7E6—Table III										14E6
14E7	Duplex-Diode Pentode	L.	8AE	14	0.16	4.6	5.3	.005	Class-A Amp.	Characteristics same as 7E7—Table III										14E7
14F7	Twin Triode	L.	8AC	14	0.16	—	—	—	Class-A Amp.	Characteristics same as 7F7—Table III										14F7
14F8	Twin Triode	L.	8BW	12.6	0.15	2.8	1.4	1.2	Class-A_1 Amp.	Characteristics same as 7F8										14F8
14H7	Semi-Variable-μ Pentode	L.	8V	14	0.16	8.0	7.0	.007	Class-A Amp.	250	– 2.5	150	3.5	9.5	800000	3800	—	—	—	14H7
14J7	Triode-Hexode Converter	L.	8BL	14	0.16	Ipt = 5 Ma.			Converter	Characteristics same as 7J7—Table III										14J7
14N7	Twin Triode	L.	8AC	14	0.32	—	—	—	Class-A Amp.	Characteristics same as 7N7—Table III										14N7
14Q7	Heptode Pentagrid Converter	L.	8AL	14	0.16	—	—	—	Converter	Characteristics same as 7Q7—Table III										14Q7
14R7	Duplex-Diode Pentode	L.	8AE	14	0.16	5.6	5.3	.004	Class-A Amp.	Characteristics same as 7R7—Table III										14R7
14S7	Triode Heptode	L.	8BL	14	0.16	Ipt = 5 Ma.			Converter	250	– 2.0	100	3	1.8	1250000	525	—	—	—	14S7
14V7	H.f. Pentode	L.	8V	14	0.24	—	—	—	Class-A Amp.	300	– 2.0	150	3.9	9.6	300000	5800	—	—	—	14V7
14W7	Pentode	L.	8BJ	14	0.24	Rk = 160 ohms			Class-A Amp.	300	– 2.2	150	3.9	10	300000	5800	—	—	—	14W7
18	Pentode	M.	6B	14	0.30	—	—	—	Class-A Amp.	Characteristics same as 6F6G										18
19BG6G	Beam Power Amp.	O.	5BT	18.9	0.3	11	6.5	0.65	Deflection Amp.	400	Peak surge E_P = 4000 V. Peak surge E_G = –100 V. I_{G2} = 6 ma. I_P = 70 ma.									19BG6G
20J8GM	Triode Heptode Converter	O.	8H	20	0.15	—	—	—	Converter	250	– 3.0	100	3.4	1.5	Triode Plate (No. 6) 100 v. 1.5 ma.					20J8GM
21A7	Triode Hexode Converter	L.	8AR	21	0.16	—	—	—	Converter	250	– 3.0	100	2.8	1.3	—	275	—	—	—	21A7
										150	– 3.0	Triode		3.5	—	1900	32	—	—	
25A6 [8]	Pentode Power Amplifier	O.	7S	25	0.3	8.5	12.5	0.20	Class-A Amp.	135	–20.0	135	8	37	35000	2450	85	4000	2.0	25A6
25A7GT [8]	Rectifier Power Pentode	O.	8F	25	0.3	—	—	—	Class-A Amp.	100	–15.0	100	4	20.5	50000	1800	90	4500	0.77	25A7GT
25AC5GT [8]	Triode Power Amplifier	O.	6Q	25	0.3	—	—	—	Class-A Amp.	110	+15.0	—	—	45	—	3800	58	2000	2.0	25AC5GT
										165	Used in dynamic-coupled circuit with 6AF5G driver							3500	3.3	
25B5 [8]	Direct-Coupled Triodes	S.	6D	25	0.3	—	—	—	Class-A Amp.	110	0	110	7	45	11400	2200	25	2000	2.0	25B5
25B6G [8]	Pentode Power Amplifier	O.	7S	25	0.3	—	—	—	Class-A Amp.	95	–15.0	95	4	45	—	4000	—	2000	1.75	25B6G
25B8GT [8]	Triode Pentode	O.	8T	25	0.15	—	—	—	Class-A Amp.		Characteristics same as 12B8GT									25B8GT
25BQ6GT	Beam Pentode	O.	6AM	25	0.3	—	—	—	Deflection Amp.	250	47*	150	2.1	45	—	5500	—	—	—	25BQ6GT
25C6G [8]	Beam Power Amplifier	O.	7AC	25	0.3	—	—	—	Class-A_1 Amp.	135	–13.5	135	3.5/11.5	58/60	9300	7000	—	2000	3.6	25C6G
25D8GT	Diode Triode Pentode	O.	8AF	25	0.15	—	—	—	Triode Amp.	100	– 1.0	—	—	0.5	91000	1100	100	—	—	25D8GT
									Pentode Amp.	100	– 3.0	100	2.7	8.5	200000	1900	—	—	—	
25L6	Beam Power Amplifier	O.	7AC	25	0.3	16	13.5	0.30	Class-A_1 Amp.	110	– 8.0	110	3.5/10.5	45/48	10000	8000	80	2000	2.2	25L6
25N6G [8]	Direct-Coupled Triodes	O.	7W	25	0.3	—	—	—	Class-A Amp.	110	0	110	7	45	11400	2200	25	2000	2.0	25N6G
26A7GT	Twin Beam-Power Audio Amplifier	O.	8BU	26.5	0.6	Each Unit Push-Pull			Class-A Amp.	26.5	– 4.5	26.5	2/5.5	20/20.5	2500	5500	—	1500	0.2	26A7GT
									Class-AB Amp. [3]	26.5	– 7.0	26.5	2/8.5	19/30	—	—	—	2500[4]	0.5	
32L7GT	Diode-Beam Tetrode	O.	8Z	32.5	0.3	—	—	—	Class-A Amp.	110	– 7.5	110	3	40	15000	6000	—	2500	1.5	32L7GT
35A5	Beam Power Amplifier	L.	6AA	35	0.15	—	—	—	Class-A_1 Amp.	110	– 7.5	110	3/7	40/41	14000	5800	—	2500	1.5	35A5
35L6G	Beam Power Amplifier	O.	7AC	35	0.15	13	9.5	0.80	Class-A_1 Amp.	110	– 7.5	110	3/7	40/41	13800	5800	—	2500	1.5	35L6G
43	Pentode Power Amplifier	M.	6B	25	0.3	8.5	12.5	0.20	Class-A Amp.	95	–15.0	95	4.0	20.0	45000	2000	90	4500	0.90	43
48 [8]	Tetrode Power Amplifier	M.	6A	30	0.4	—	—	—	Class-A Amp.	96	–19.0	96	9.0	52.0	—	3800	—	1500	2.0	48
50A5	Beam Power Amplifier	L.	6AA	50	0.15	—	—	—	Class-A_1 Amp.	110	– 7.5	110	4/11	49/50	10000	8200	—	2000	2.2	50A5
50C6GT	Beam Power Amplifier	O.	7AC	50	0.15	—	—	—	Class-A_1 Amp.	135	–13.5	135	3.5/11.5	58/60	9300	7000	—	2000	3.6	50C6GT
50L6GT	Beam Power Amplifier	O.	7AC	50	0.15	—	—	—	Class-A Amp.	110	– 7.5	110	4/11	49/50	—	8200	82	2000	2.2	50L6GT
70A7GT	Diode-Beam Tetrode	O.	8AB[1]	70	0.15	—	—	—	Class-A Amp.	110	– 7.5	110	3.0	40	—	5800	80	2500	1.5	70A7GT
70L7GT	Diode-Beam Tetrode	O.	8AA	70	0.15	—	—	—	Class-A_1 Amp.	110	– 7.5	110	3/6	40/43	15000	7500	—	2000	1.8	70L7GT
117L7GT/117M7GT	Rectifier-Amplifier	O.	8AO	117	0.09	—	—	—	Class-A Amp.	105	– 5.2	105	4/5.5	43	17000	5300	—	4000	0.85	117L7GT/117M7GT
117N7GT	Rectifier-Amplifier	O.	8AV	117	0.09	—	—	—	Class-A Amp.	100	– 6.0	100	5.0	51	16000	7000	—	3000	1.2	117N7GT
117P7GT	Rectifier-Amplifier	O.	8AV	117	0.09	—	—	—	Class-A Amp.	105	– 5.2	105	4/5.5	43	17000	5300	—	4000	0.85	117P7GT

TABLE IX—HIGH-VOLTAGE HEATER TUBES—*Continued*

Type	Name	Base	Socket Connections	Heater Volts	Heater Amp.	Capacitance $\mu\mu$fd. In	Capacitance $\mu\mu$fd. Out	Capacitance $\mu\mu$fd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
1280	Pentode	L.	8V	12.6	0.15	6.0	6.5	0.007	Class-A_1 Amp.	Same as 14C7 (Special Non-microphonic)										1280
1284	U.h.f. Pentode	L.	8V	12.6	0.15	5.0	6.0	0.01	Class-A Amp.	250	− 3.0	100	2.5	9.0	800000	2000	—	—	—	1284
1629	Electron-Ray Tube	O.	6RA	12.6	0.15	—	—	—	Indicator Tube	Characteristics same as 6E5—Table IV										1629
1631	Beam Power Amplifier	O.	7AC	12.6	0.45	—	—	—	Class-A Amp.	Characteristics same as 6L6—Table I										1631
1632	Beam Power Amplifier	O.	7AC	12.6	0.6	—	—	—	Class-A Amp.	Characteristics same as 25L6										1632
1633	Twin Triode	O.	8BD	25	0.15	—	—	—	Class-A Amp.	Characteristics same as 6SN7GT—Table I										1633
1634	Twin Triode	O.	8S	12.6	0.15	—	—	—	Class-A Amp.	Characteristics same as 6SC7—Table I										1634
1644	Twin Pentode	O.	Fig. 7	12.6	0.15	—	—	—	Class-A Amp.	180	− 9.0	180	2.8/4.6	13	160000	2150	—	10000	1.0	1644
XXD/14AF7	Twin Triode	L.	8AC	12.6	0.15	—	—	—	Class-A Amp.	250	−10	—	—	9.0	—	2100	16	—	—	XXD/14AF7
28D7	Double Beam Power Amplifier	L.	8BS	28.0	0.4	—	—	—	Class-A Amp.	28	390* 180*	28 [2] 28 [3]	0.7 [2] 1.2 [3]	9.0 [2] 18.5 [3]	— —	— —	— —	4000 [4] 6000 [4]	0.08 [2] 0.175 [3]	28D7

* Cathode resistor—ohms.

[1] 6.3-volt pilot lamp must be connected between Pins 6 and 7.
[2] Per section—resistance-coupled.
[3] P.p. operation—values for both sections.
[4] Plate to plate.
[5] Values are for each unit.
[6] Values are for single tube.
[7] Grids 2 and 3 connected to plate.
[8] Discontinued.

TABLE X—SPECIAL RECEIVING TUBES

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance $\mu\mu$fd. In	Capacitance $\mu\mu$fd. Out	Capacitance $\mu\mu$fd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
00-A [7]	Triode Detector	M.	4D	5.0	0.25	3.2	2.0	8.50	Grid-Leak Det.	45	—	—	—	1.5	30000	666	20	—	—	00-A
01-A [7]	Triode Detector Amplifier	M.	4D	5.0	0.25	—	—	—	Class-A Amp.	135	− 9.0	—	—	3.0	10000	800	8.0	—	—	01-A
3A8GT	Diode Triode Pentode	O.	8AS	1.4	0.1	2.6	4.2	2.0	Class-A Triode	90	0	—	—	0.15	240000	275	65	—	—	3A8GT
				2.8	0.05	3.0	10.0	0.012	Class-A Pentode	90	0	90	0.3	1.2	600000	750	—	—	—	
3B5GT	Beam Power Amplifier	O.	7AP	1.4 2.8	0.1 0.05	—	—	—	Class-A Amp.	67.5	− 7.0	67.5	0.6 0.5	8.0 6.7	100000	1650 1500	—	5000	0.2 0.18	3B5GT
3C5GT	Power Output Pentode	O.	7AQ	1.4 2.8	0.1 0.05	—	—	—	Class-A Amp.	90	− 9.0	90	1.4	6.0	—	1550 1450	—	8000 10000	0.24 0.26	3C5GT
3C6	Twin Triode	L.	7BW	1.4 2.8	0.1 0.05	—	—	—	Class-A Amp.	90	0	—	—	4.5	11200	1300	14.5	—	—	3C6
3LE4	Power Amplifier Pentode	L.	6BA	2.8	0.05	—	—	—	Class-A Amp.	90	− 9.0	90	1.8	9.0	110000	1600	—	6000	0.30	3LE4
3LF4	Power Amplifier Tetrode	L.	6BB	1.4 2.8	0.1 0.05	—	—	—	Class-A Amp.	90	− 4.5	90	1.3 1.0	9.5 8.0	75000 80000	2200 2000	—	8000 7000	0.27 0.23	3LF4
3Q5GT	Beam Power Amplifier	O.	7AQ	1.4 2.8	0.1 0.05	Parallel Filaments Series Filaments			Class-A Amp.	90	− 4.5	90	1.3 1.0	9.5 7.5	—	2100 1800	—	8000	0.27 0.25	3Q5GT
4A6G	Twin Triode Amplifier	O.	8L	4	0.06	Triodes Parallel			Class-A Amp.	90	− 1.5	—	—	2.2	13300	1500	20	—	—	4A6G
				2	0.12	Both Sections			Class-B Amp.	90	0	—	—	4.6	—	—	—	8000	1.0	
6F4	Acorn Triode	A.	7BR	6.3	0.225	2.0	0.6	1.90	Class-A Amp.	80	150*	—	—	13.0	2900	5800	17	—	—	6F4
6L4	U.H.F. Triode	A.	7BR	6.3	0.225	1.8	0.5	1.6	Class-A_1 Amp.	80	150*	—	—	9.5	4400	6400	28	—	—	6L4
10	Triode Power Amplifier	M.	4D	7.5	1.25	4.0	3.0	7.00	Class-A Amp.	425	−39.0	—	—	18.0	5000	1600	8.0	10200	1.6	10
11/12 [7]	Triode Detector Amplifier	M.	4F/4D	1.1	0.25	—	—	—	Class-A Amp.	135	−10.5	—	—	3.0	15000	440	6.6	—	—	11/12
20 [7]	Triode Power Amplifier	S.	4D	3.3	0.132	2.0	2.3	4.10	Class-A Amp.	135	−22.5	—	—	6.5	6300	525	3.3	6500	0.11	20
22 [7]	Tetrode R.F. Amplifier	M.	4K	3.3	0.132	3.5	10	0.02	Class-A Amp.	135	− 1.5	67.5	1.3	3.7	325000	500	160	—	—	22
26	Triode Amplifier	M.	4D	1.5	1.05	2.8	2.5	8.10	Class-A Amp.	180	−14.5	—	—	6.2	7300	1150	8.3	—	—	26
40 [7]	Triode Voltage Amplifier	M.	4D	5.0	0.25	2.8	2.2	2.00	Class-A Amp.	180	− 3.0	—	—	0.2	150000	200	30	—	—	40
50	Triode Power Amplifier	M.	4D	7.5	1.25	4.2	3.4	7.10	Class-A Amp.	450	−84.0	—	—	55.0	1800	2100	3.8	4350	4.6	50

TABLE X—SPECIAL RECEIVING TUBES—*Continued*

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
71-A	Triode Power Amplifier	M.	4D	5.0	0.25	3.2	2.9	7.50	Class-A Amp.	180	−43.0	—	—	20.0	1750	1700	3.0	4800	0.79	71-A
99 [6]	Triode Detector Amplifier	S.	4D	3.3	0.063	2.5	2.5	3.30	Class-A Amp.	90	− 4.5	—	—	2.5	15500	425	6.6	—	—	99
112A [7]	Triode Detector Amplifier	M.	4D	5.0	0.25	—	—	—	Class-A Amp.	180	−13.5	—	—	7.7	4700	1800	8.5	—	—	112A
182B/ 482B	Triode Amplifier	M.	4D	5.0	1.25	—	—	—	Class-A Amp.	250	−35.0	—	—	18.0	—	1500	5.0	—	—	182B/ 482B
183/483 [7]	Power Triode	M.	4D	5.0	1.25	—	—	—	Class-A Amp.	250	−60.0	—	—	25.0	18000	1800	3.2	4500	2.0	183/483
485 [7]	Triode	S.	5A	3.0	1.3	—	—	—	Class-A Amp.	180	− 9.0	—	—	6.0	9300	1350	12.5	—	—	485
864	Triode Amplifier	S.	4D	1.1	0.25	—	—	—	Class-A Amp.	90	− 4.5	—	—	2.9	13500	610	8.2	—	—	864
954	Pentode Detector, Amplifier	A.	5BB	6.3	0.15	3.4	3.0	0.007	Class-A Amp.	250	− 3.0	100	0.7	2.0	1.5 meg.	1400	2000	—	—	954
									Bias Detector	250	− 6.0	100	—	Plate current to be adjusted to 0.1 ma. with no signal					—	
955	Triode Detector, Amplifier, Oscillator	A.	5BC	6.3	0.15	1.0	0.6	1.40	Class-A Amp.	250	− 7.0	—	—	6.3	11400	2200	25	—	—	955
										90	− 2.5	—	—	2.5	14700	1700	25	—	—	
956	Variable-μ Pentode R.F. Amplifier	A.	5BB	6.3	0.15	3.4	3.0	0.007	Class-A Amp.	250	− 3.0	100	2.7	6.7	700000	1800	1440	—	—	956
									Mixer	250	−10.0	100	—	—	Oscillator peak volts—7 min.					
957	Triode Detector, Amplifier, Oscillator	A.	5BD	1.25	0.05	0.3	0.7	1.20	Class-A Amp.	135	− 5.0	—	—	2.0	20800	650	13.5	—	—	957
958 958-A	Triode A.F. Amplifier, Oscillator	A.	5BD	1.25	0.1	0.6	0.8	2.60	Class-A Amp.	135	− 7.5	—	—	3.0	10000	1200	12	—	—	958 958-A
959	Pentode Detector, Amplifier	A.	5BE	1.25	0.05	1.8	2.5	0.015	Class-A Amp.	145	− 3.0	67.5	0.4	1.7	800000	600	480	—	—	959
7E5/1201	U.h.f. Triode	L.	8BN	6.3	0.15	3.6	2.8	1.50	Class-A Amp.	180	− 3	—	—	5.5	12000	—	36	—	—	7E5/1201
7C4/1203	U.h.f. Diode	L.	4AH	6.3	0.15	—	—	—	Rectifier		Max. r.m.s. voltage—150				Max. d.c. output current—8 ma.					7C4/1203
7AB7/ 1204	Sharp Cut-off Pentode	L.	8BO	6.3	0.15	3.5	4.0	0.06	Class-A Amp.	250	− 2	100	0.6	1.75	800000	1200	—	—	—	7AB7/ 1204
1276	Triode Power Amplifier	M.	4D	4.5	1.14	—	—	—	Class-A Amp.				Characteristics similar to 6A3							1276
1609	Pentode Amplifier	S.	5B	1.1	0.25	—	—	—	Class-A Amp.	135	− 1.5	67.5	0.65	2.5	400000	725	300	—	—	1609
9004	U.h.f. Diode	A.	4BJ	6.3	0.15	—	—	—	Detector			Max. a.c. voltage—117. Max. d.c. output current—5 ma.								9004
9005	U.h.f. Diode	A.	5BG	3.6	0.165	—	—	—	Detector			Max. a.c. voltage—117. Max. d.c. output current—1 ma.								9005
EF-50	Sharp Cut-off Pentode	L.	9C	6.3	0.3	8	5	0.007	I.F.-R.F. Amp.	250	150*	250	3.1	10	600000	6300	—	—	—	EF-50
GL-2C44 GL-464A	U.h.f. Triode	O.	Fig. 17	6.3	0.75	—	—	—	Class-A Amp. and Modulator	250	100*	—	—	25.0	—	7000	—	—	—	GL-2C44 GL-464A
GL-446A GL-446B	U.h.f. Triode	O.	Fig. 19	6.3	0.75	—	—	—	Oscillator, Amp. or Converter	250	200*	—	—	15.0	—	4500	45	—	—	GL-446A GL-446B
559 GL-559	U.h.f. Diode	O.	Fig. 18	6.3	0.75	—	—	—	Detector or trans. line switch	5.0	—	—	—	24.0	—	—	—	—	—	559 GL-559
NU-2C35	Special Hi-Mu Triode	O.	Fig. 38	6.3	0.3	5.2	2.3	0.62	Shunt Voltage Regulator	8000	−200	—	—	5.0	525000	950	500	—	—	NU-2C35
VT52	Triode	M.	4D	7.0	1.18	5.0	3.0	7.7	Class-A_1 Amp.	220	−43.5	—	—	29.0	1650	2300	3.8	3800	1.0	VT52
X6030	Diode	L.	Fig. 4	3.0	0.6	—	—	—	Noise Diode	90	—	—	—	4.0	—	—	—	—	—	X6030

TABLE X—SPECIAL RECEIVING TUBES—*Continued*

Type	Name	Base	Socket Connec-tions	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transcon-ductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
XXB	Twin-Triode Frequency Converter	L.	Fig. 9	2.8/ 1.4 3.2[3]/ 1.6	0.05/ 0.10 — —	—	—	—	Converter[2]	90[1]	0	—	—	4.5[4] 4.5[5]	11200[4] 11200[5]	1300[4] 1300[5]	14.5[1]	—	—	XXB
											−3	—	—	1.4[4] 1.4[5]	1900[4] 1900[5]	760[4] 760[5]	14.5[1]	—	—	
XXFM	Twin-Diode Triode	L.	8BZ	6.3	0.3	—	—	—	Class-A Amp.	250	−1	—	—	1.9	6700	1500	100	—	—	XXFM
										100	0	—	—	1.2	85000	1000	85	—	—	

* Cathode resistor—ohms.
[1] Both sections.
[2] Section No. 2 recommended for h.f.o.
[3] Dry battery operation.
[4] Section No. 1.
[5] Section No. 2.
[6] Same as X99. Type V99 is same, but socket connections are 4E.
[7] Discontinued.

TABLE XI—MINIATURE RECEIVING TUBES

Other miniature types in Tables XIII and XV

Type	Name	Base	Socket Connec-tions	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transcon-ductance Micromhos	Amp. Factor [4]	Load Resistance Ohms	Power Output Watts	Prototype
1A3	H. F. Diode	B.	5AP	1.4	0.15	—	—	—	Detector F.M. Discrim.		Max. a.c. voltage per plate—117. Max. output current—0.5 ma.									—
1L4	Sharp Cut-off Pentode	B.	6AR	1.4	0.05	3.6	7.5	.008	Class-A Amp.	90	0	90	2.0	4.5	350000	1025	—	—	—	1N5GT
1R5	Pentagrid Converter	B.	7AT	1.4	0.05	—	—	—	Converter	90	0	67.5	3.0	1.7	500000	300	Grid No. 1 100000 ohms			1A7GT
1S4	Pentagrid Power Amp.	B.	7AV	1.4	0.1	—	—	—	Class-A Amp.	90	−7.0	67.5	1.4	7.4	100000	1575	—	8000	0.270	1Q5GT
1S5	Diode Pentode	B.	6AU	1.4	0.05	—	—	—	Class-A Amp.	67.5	0	67.5	0.4	1.6	600000	625	—	—	—	—
									R-Coupled Amp.	90	0	90	Screen resistor 3 meg., grid 10 meg.					1 meg.	0.050	
1T4	Variable-μ Pentode	B.	6AR	1.4	0.05	3.6	7.5	0.01	Class-A Amp.	90	0	67.5	1.4	3.5	500000	900	—	—	—	1P5GT
1U4	Sharp Cut-off Pentode	B.	6AR	1.4	0.05	3.6	7.5	0.01	Class-A Amp.	90	0	90	0.5	1.6	1500000	900	—	—	—	1N5GT
1U5	Diode Pentode	B.	6BW	1.4	0.05	—	—	—	Class-A Amp.	67.5	0	67.5	0.4	1.6	600000	625	—	—	—	—
2C51	Twin Triode	B.	8CJ	6.3	0.3	2.2	1.0	1.3	Class-A₁ Amp.	150	−2	—	—	8.2[1]	—	5500	35	—	—	7F8
2E30	Beam Power Pentode	B.	7CQ	6.0	0.7	10	4.5	0.5	Class-A₁ Single	250	450*	250	7.4[2]	44[2]	63000	3700	40[5]	4500	4.5	—
									Class-A₁ Amp.[3]	250	225*	250	14.8[2]	88[2]	—	—	80[5]	9000[6]	9	
									Class-AB₁ Amp.[3]	250	−25	250	13.5[2]	80[2]	—	—	48[5]	8000[6]	12.5	
									Class-AB₂ Amp.[3]	250	−30	250	20[2]	120[2]	—	—	40[5]	3800[6]	17	
3A4	Power Amplifier Pentode	B.	7BB	1.4 2.8	0.2 0.1	4.8	4.2	0.34	Class-A₁ Amp.	135 150	−7.5 −8.4	90 90	2.6 2.2	14.9[2] 14.1[2]	90000 100000	1900	—	8000	0.6 0.7	—
3A5	H.F. Twin Triode	B.	7BC	1.4 2.8	0.22 0.11	0.9	1.0	3.20	Class-A Amp.	90	−2.5	—	—	3.7	8300	1800	15	—	—	—
3Q4	Power Amplifier Pentode	B.	7BA	1.4 2.8	0.1 0.05	Parallel Filaments Series Filaments			Class-A Amp.	90	−4.5	90	2.1 1.7	9.5 7.7	100000 120000	2150 2000	—	10000	0.27 0.24	3Q5GT
3S4	Power Amplifier Pentode	B.	7BA	1.4 2.8	0.1 0.05	Parallel Filaments Series Filaments			Class-A Amp.	90	−7.0	67.5	1.4 1.1	7.4 6.1	100000	1575 1425	—	8000	0.27 0.235	3Q5GT
3V4	Power Amplifier Pentode	B.	6BX	1.4	0.1	Parallel Filaments			Class-A Amp.	90	−4.5	90	2.1	9.5	100000	2150	—	10000	0.27	3Q5GT
				2.8	0.05	Series Filaments			Class-A Amp.	90	−4.5	90	1.7	7.7	120000	2000	—	10000	0.24	
6AB4	Triode R.F. Amp.	B.	5CE	6.3	0.15	2.2	0.5	1.5	Class-A Amp.	250	−2	—	—	10	—	5500	55	—	—	Single unit 12AT7
6AG5	Sharp Cut-off Pentode	B.	7BD	6.3	0.3	—	—	—	Class-A Amp.	250 100	200* 100*	150 100	2.0 1.6	7.0 5.5	800000 300000	5000 4750	— —	— —	—	6SH7GT
6AH6	Sharp Cut-off Pentode	B.	7CC	6.3	0.45	10	2	0.03	Pentode Amp.	300	160*	150	2.5	10	500000	9000	—	—	—	6AC7
									Triode Amp.[5]	150	160*	—	—	12.5	3600	11000	40	—	—	
6AJ5	Sharp Cut-off Pentode	B.	7PM	6.3	0.175	—	—	—	R.F. Amplifier	28	200*	28	1.2	3.0	90000	2750	250	—	—	—
									Class-AB Amp.[3]	180	−7.5	75	—	—	—	—	—	28000[6]	1.0	
6AK5	Sharp Cut-off Pentode	B.	7BD	6.3	0.175	4.3	2.1	0.03	R.F. Amplifier	180	200*	120	2.4	7.7	690000	5100	3500	—	—	—
										150	330*	140	2.2	7.0	420000	4300	1800	—	—	
										120	200*	120	2.5	7.5	340000	5000	1700	—	—	

TABLE XI—MINIATURE RECEIVING TUBES—Continued

Type	Name	Base	Socket Connections[1]	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor [4]	Load Resistance Ohms	Power Output Watts	Prototype
6AK6	Power Amplifier Pentode	B.	7BK	6.3	0.15	3.6	4.2	0.12	Class-A Amp.	180	− 9.0	180	2.5	15.0	200000	2300	—	10000	1.1	—
6AL5	U.h.f. Twin Diode	B.	6BT	6.3	0.3	—	—	—	Detector			Max. r.m.s. voltage—150. Max. d.c. output current—10 ma.[1]								6H6GT
6AN5	Power Amp. Pentode	B.	7BD	6.3	0.5	9.0	4.8	0.05	Class-A1 Amp.	120	− 6	120	12	35	12500	8000	—	—	—	6AG7
6AN6	Twin Diode	B.	7BJ	6.3	0.2	—	—	—	Detector	R.m.s. voltage per plate = 75 volts; d.c. output = 3.5 ma. with 25000 ohms and 8 μμfd. load; peak current per plate = 10 ma.; peak inverse voltage = 210.										—
6AQ5	Beam Power Tetrode	B.	7BZ	6.3	0.45	7.6	6.0	0.35	Class-A1 Amp.	180	− 8.5	180	4.0 [2]	30 [2]	58000	3700	29 [5]	5500	2.0	6V6GT
										250	−12.5	250	7.0 [2]	47 [2]	52000	4100	45 [5]	5000	4.5	
6AQ6	Duodiode Hi-mu Triode	B.	7BT	6.3	0.15	1.7	1.5	1.80	Class-A Triode	250	− 3.0	—	—	1.0	58000	1200	70	—	—	6T7G
										100	− 1.0	—	—	0.8	61000	1150	70	—	—	
6AR5	Pentode Power Amp.	B.	6CC	6.3	0.4	—	—	—	Class-A1 Amp.	250	−18	250	5.5 [2]	33 [2]	68000	2300	—	7600	3.4	6K6GT
										250	−16.5	250	5.5 [2]	35 [2]	65000	2400	—	7000	3.2	
6AS5	Beam Pentode	B.	7CV	6.3	0.8	12	6.2	0.6	Class-A1 Amp.	150	− 8.5	110	2/6.5	35/36	—	5600		4500	2.2	—
6AS6	Sharp Cut-off Pentode	B.	7CM	6.3	0.175	4.0	3.0	0.02	Class-A Amp.	120	− 2	120	3.5	5.5	—	3500	—	—	—	—
6AT6	Duplex Diode Triode	B.	7BT	6.3	0.3	2.3	1.1	2.10	Class-A Amp.	250	− 3	—	—	1.0	58000	1200	70	—	—	6Q7GT
6AU6	Sharp Cut-off Pentode	B.	7BK	6.3	0.3	5.5	5.0	.0035	Class-A Amp.	250	− 1	150	4.3	10.8	2000000	5200	—	—	—	6SH7GT
6AV6	Duodiode Hi-mu Triode	B.	7BT	6.3	0.3	—	—	—	Class-A1 Amp.	250	− 2	—	—	1.2	62500	1600	100	—	—	6SQ7GT
6BA6	Remote Cut-off Pentode	B.	7CC	6.3	0.3	5.5	5.0	.0035	Class-A Amp.	250	68*	100	4.2	11	1500000	4400	—	—	—	6SG7GT
6BA7	Pentagrid Converter	B.	8CT	6.3	0.3	9.5	8.3	—	Converter	250	− 1	100	10	3.8	1000000	3.5	—	—	—	6SB7Y
6BD6	Remote Cut-off Pentode	B.	7CC	6.3	0.3	—	—	—	Class-A Amp.	100	− 1	100	5	13	120000	2350	—	—	—	6SK7GT
										250	− 3	100	3.5	9	700000	2000	—	—	—	
6BE6	Pentagrid Converter	B.	7CH	6.3	0.3	Osc. Grid 50000 Ω			Converter	250	− 1.5	100	7.8	3.0	1000000	475	—	—	—	6SA7GT
6BF6	Duplex-Diode Triode	B.	7BT	6.3	0.3	1.8	1.1	2.0	Class-A1 Amp.	250	− 9	—	—	9.5	8500	1900	16	10000	—	6SR7GT
6BH6	Sharp Cut-off Pentode	B.	7CM	6.3	0.15	5.4	4.4	0.0035	Class-A1 Amp.	250	− 1	150	2.9	7.4	1400000	4600	—	—	—	—
6BJ6	Remote Cut-off Pentode	B.	7CM	6.3	0.15	4.5	5.0	.0035	Class-A1 Amp.	250	− 1	100	3.3	9.2	1300000	3800	—	—	—	6SS7GT
6C4	Triode Amplifier	B.	6BG	6.3	0.15	1.8	1.3	1.60	Class-A1 Amp.	250	− 8.5	—	—	10.5	7700	2200	17	—	—	6J5GT
6J4	U.h.f. Grounded-Grid R.F. Amplifier	B.	7BQ	6.3	0.4	5.5	0.24	4.0	Grounded-Grid	150	200*	—	—	15.0	4500	12000	55	—	—	—
									Class-A1 Amp.	100	100*	—	—	10.0	5000	11000	55	—	—	
6J6	Twin Triode	B.	7BF	6.3	0.45	2.2	0.4	1.6	Class-A1 Amp. Mixer, Oscillator	100	50*	—	—	8.5	7100	5300	38	—	—	—
6N4	U.h.f. Triode Amplifier	B.	7CA	6.3	0.2	3.0	1.6	1.10	Class-A Amp.	180	− 3.5	—	—	12.0	—	6000	32	—	—	—
6T8	Triple-Diode Triode	B.	9E	6.3	0.45	1.5	1.1	2.4	Class-A1 Amp.	250	− 3	—	—	1.0	5800	1200	70	—	—	—
										100	− 1	—	—	0.8	5400	1300	70	—	—	
12AL5	Twin Diode	B.	6BT	12.6	0.15	2.5	—	—	Detector		R.m.s. voltage per plate = 117; d.c. output = 9 ma. per plate; peak ma. per plate = 54; peak inverse voltage = 330.									12H6GT
12AT6	Duplex Diode Triode	B.	7BT	12.6	0.15	2.3	1.1	2.10	Class-A Amp.	250	− 3.0	—	—	1.0	58000	1200	70	—	—	12Q7GT
12AT7	Double Triode	B.	9A	6.3	0.3	2.5 [7]	0.45 [7]	1.45 [7]	Class-A1 Amp.	250	− 2	—	—	10	10000	5500	55	—	—	—
				12.6	0.15	2.5 [8]	0.35 [8]	1.45 [8]	Each Unit	180	− 1	—	—	11	9400	6600	62	—	—	
12AU6	Sharp Cut-off Pentode	B.	7CC	12.6	0.15	5.5	5.0	.0035	Class-A1 Amp.	250	− 1.0	150	4.3	10.8	1 meg.	5200	—	—	—	12SH7GT
12AU7	Twin-Triode Amplifier	B.	9A	6.3	0.3	1.6 [7]	0.5 [7]	1.5 [7]	Class-A1 Amp.	250	− 8.5	—	—	10.5	7700	2200	17	—	—	12SN7GT
				12.6	0.15	1.6 [8]	0.35 [8]	1.5 [8]												
12AV6	Duodiode Hi-mu Triode	B.	7BT	12.6	0.15	—	—	—	Class-A1 Amp.	250	− 2	—	—	1.2	62500	1600	100	—	—	—
12AW6	Sharp Cut-off Pentode	B.	7CM	12.6	0.15	6.5	1.5	0.025	Pentode Amp.	250	200*	150	2.0	7.0	800000	5000	—	—	—	—
									Triode Amp. [9]	250	825*	—	—	5.5	11000	3800	42	—	—	
12AW7	Sharp Cut-off Pentode	B.	7CM	12.6	0.15	6.5	1.5	0.025	Class-A1 Amp.	250	200*	150	2.0	7.0	0.8 meg.	5000	—	—	—	—
12AX7	Double Triode	B.	9A	12.6	0.15	1.6 [7]	0.46 [7]	1.7 [7]	Class-A1 Amp.	250	− 2	—	—	1.2 [1]	62500	1600	100	—	—	—
				6.3	0.3	1.6 [8]	0.34 [8]	1.7 [8]		100	− 1	—	—	0.5 [1]	8000	1250	100	—	—	
12AY7	Dual Triode	B.	9A	12.6 6.3	0.15 0.3	1.3	0.6	1.3	Class-A Amp.	250	− 4	—	—	3	—	1750	40	—	—	—
									Lo-Level Amp.	150	2700*		Plata resistor = 20000 Ω. Grid resistor = 0.1 Meg. V.G. = 12.5							
12BA6	Remote Cut-off Pentode	B.	7CC	12.6	0.15	5.5	5.0	.0035	Class-A Amp.	250	68*	100	4.2	11.0	1500000	4400	—	—	—	12SG7G

TABLE XI — MINIATURE RECEIVING TUBES — *Continued*

Type	Name	Base	Socket Connections[1]	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor[4]	Load Resistance Ohms	Power Output Watts	Prototype
12BA7	Pentagrid Converter	B.	8CT	12.6	0.15	9.5	8.3	—	Converter	250	− 1	100	10	3.8	1000000	3.5	—	—	—	—
12BD6	Remote Cut-off Pentode	B.	7CC	12.6	0.15	4.3	5.0	.004	Class-A Amp.	250	− 3	100	3.5	9.0	700000	2000	—	—	—	12SK7GT
12BE6	Pentagrid Converter	B.	7CH	12.6	0.15	Osc. Grid 50000 Ω			Converter	250	− 1.5	100	7.8	3.0	1000000	475	—	—	—	12SA7GT
12BF6	Duodiode Triode	B.	7BT	12.6	0.15	1.8	1.1	2.00	Class-A Amp.	250	− 9	—	—	9.5	8500	1900	16	—	—	12SR7GT
19J6	Twin Triode	B.	7BF	18.9	0.15	2.0	0.4	1.5	Class-A_1 Amp.	100	50*	—	—	8.5 [1]	7100	5300	38	—	—	—
19T8	Triple-Diode Triode	B.	9E	18.9	0.15	1.5	1.1	2.4	Class-A_1 Amp.	250	− 3	—	—	1.0	5800	1200	70	—	—	—
26A6	Remote Cut-off Pentode	B.	7BK	26.5	0.07	6.0	5.0	.0035	Class-A_1 Amp.	250	125*	100	4	10.5	1000000	4000	—	—	—	—
26C6	Duplex-Diode Triode	B.	7BT	26.5	0.07	1.8	1.4	2	Class-A_1 Amp.	250	− 9	—	—	9.5	8500	1900	16	—	—	—
26D6	Pentagrid Converter	B.	7CH	26.5	0.07	Osc. Grid 20000 Ω			Converter	250	− 1.5	100	7.8	3.0	1000000	475	—	—	—	—
35B5	Beam Power Amplifier	B.	7BZ	35	0.15	11	6.5	0.4	Class-A_1 Amp.	110	− 7.5	110	7 [2]	41 [2]	—	5800	40 [5]	2500	1.5	35L6GT
35C5	Beam Power Amplifier	B.	7CV	35	0.15	12	6.2	0.57	Class-A_1 Amp.	110	− 7.5	110	3/7	40/41	—	5800		2500	1.5	—
50B5	Beam Power Amplifier	B.	7BZ	50	0.15	13	6.5	0.50	Class-A Amp.	110	− 7.5	110	4.0	49.0	14000	7500	—	3000	1.9	50L6GT
50C5	Beam Power Amplifier	B.	7CV	50	0.15	—	—	—	Class-A_1 Amp.	110	− 7.5	110	4/8.5	49/50	10000	7500	—	2500	1.9	—
5590	Pentode	B.	7BD	6.3	0.15	3.4	2.9	0.01	Class-A_1 Amp.	90	820*	90	1.4	3.9	300000	2000	—	—	—	—
5591	R.F. Pentode	B.	7BD	6.3	0.15	3.9	2.85	0.01	Class-A_1 Amp.	180	200*	120	2.4	1.7	690000	5100	3500	—	—	—
5654	Sharp Cut-off Pentode	B.	7BD	6.3	0.175	4	2.9	0.02	Class-A_1 Amp.	120	200*	120	2.5	7.5	340000	5000	—	—	—	—
5687	Dual Triode	B.	9H	12.6 / 6.3	0.45 / 0.9	4	0.45	3.1	Class-A Amp.	250	−12.5	—	—	16	4000	4100	16.5	—	—	—
										120	− 2	—	—	34	2000	10000	20	—	—	
5722	Noise Generating Diode	B.	5CB	2/5.5	1.6	—	1.5	—	Noise Generator	200	—	—	—	35	—	—	—	—	—	—
9001	Sharp Cut-off Pentode	B.	7PM	6.3	0.15	3.6	3.0	0.01	Class-A Amp.	250	− 3.0	100	0.7	2.0	1 meg. +	1400	—	—	—	—
									Mixer	250	− 5.0	100	Osc. peak voltage 4 volts			550	—	—	—	
9002	Triode Detector, Amplifier, Oscillator	B.	7TM	6.3	0.15	1.2	1.1	1.40	Class-A Amp.	250	− 7.0	—	—	6.3	11400	2200	25	—	—	—
										90	− 2.5	—	—	2.5	14700	1700	25	—	—	
9003	Remote Cut-off Pentode	B.	7PM	6.3	0.15	3.6	3.0	0.01	Class-A Amp.	250	− 3.0	100	2.7	6.7	700000	1800	—	—	—	—
									Mixer	250	−10.0	100	Osc. peak voltage 9 volts			600	—	—	—	
9006	U.h.f. Diode	B.	6BH	6.3	0.15	—	—	—	Detector	Max. a.c. voltage—270. Max. d.c. output current—5 ma.										—

* Cathode resistor—ohms.

[1] Per Plate.
[2] Maximum-signal current for full-power output.
[3] Values are for two tubes in push-pull.
[4] Also no-signal plate ma. when so indicated.
[5] No signal plate ma.
[6] Effective plate-to-plate.
[7] Triode No. 1.
[8] Triode No. 2.
[9] Grid No. 2 tied to plate and No. 3 to cathode.

TABLE XII — SUB-MINIATURE TUBES

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
1AC5	Power Pentode	Bs.	Fig. 14	1.25	0.04	—	—	—	Class-A_1 Amp.	67.5	−4.5	67.5	0.4	2.0	150000	750	—	25000	0.05	1AC5
1AD5	Sharp Cut-off Pentode	Bs.	Fig. 16	1.25	0.04	1.8	2.8	0.01	Class-A_1 Amp.	67.5	0	67.5	0.75	1.85	700000	735	—	—	—	1AD5
1C8	Heptode	—	—	1.25	0.04	6.5	4.0	0.25	Converter	30	0	30	0.75	0.32	300000	100	—	—	—	1C8
1E8	Pentagrid Convertor	Bs.	Fig. 27	1.25	0.04	6	—	—	Converter	67.5	0	67.5	1.5	1.0	—	150	—	—	—	1E8
1T6	Diode-Pentode	Bs.	Fig. 28	1.25	0.04	—	—	—	Class-A_1 Amp.	67.5	0	67.5	0.4	1.6	400000	600	—	—	—	1T6
1V5	Audio Pentode	[1]	[2]	1.25	0.04	—	—	—	Class-A_1 Amp.	67.5	−4.5	67.5	0.4	2.0	150000	750	—	25000	0.05	1V5
1W5	Sharp Cut-off Pentode	[1]	[2]	1.25	0.04	2.3	3.5	0.01	Class-A_1 Amp.	67.5	0	67.5	0.75	1.85	700000	735	—	—	—	1W5
2E31	R.F. Pentode	[1]	[2]	1.25	0.05				Class-A_1 Amp.	22.5	0	22.5	0.3	0.4	—	500	—	—	—	2E31
2E32	R.F. Pentode	[1]	[2]	1.25	0.05				Class-A Amp.	22.5	0	22.5	0.3	0.4	350000	500	—	—	—	2E32
2E35	Audio Pentode	[1]	[2]	1.25	0.03				Class-A_1 Amp.	22.5	0	22.5	0.07	0.27	—	385	—	—	0.0012	2E35
2E36	Audio Pentode	[1]	[3]	1.25	0.03				Class-A_1 Amp.	22.5	0	22.5	0.07	0.27	220000	385	—	150000	0.0012	2E36
										45	−1.25	45	0.11	0.45	250000	500	—	100000	0.006	
2E41	Diode Pentode	[1]	[3]	1.25	0.03	—	—	—	Detector Amp.	22.5	0	22.5	0.12	0.35	—	—	—	—	—	2E41
2E42	Diode Pentode	[1]	[3]	1.25	0.03				Detector Amp.	22.5	0	22.5	0.12	0.35	250000	375	—	1 meg.	—	2E42

TABLE XII — SUB-MINIATURE TUBES — *Continued*

Type	Name	Base	Socket Connec-tions	Fil. or Heater		Capacitance μμfd.			Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transcon-ductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
				Volts	Amp.	In	Out	Plate-Grid												
2G21	Triode Heptode	[1]	[2]	1.25	0.05	—	—	—	Converter	22.5	—	22.5	0.2	0.3	—	75	—	—		2G21
2G22	Converter	[1]	[2]	1.25	0.05				Converter	22.5	0	22.5	0.3	0.2	500000	60	—	—	—	2G22
6K4	Triode	[1]	[2]	6.3	0.15	2.4	0.8	2.4	Class A_1 Amp.	200	680*	—	—	11.5	4650	3450	16	—	—	6K4
1247	Diode	[1]	[2]	0.7	0.065	—	—	—	R.F. Probe			Max. a.c. volts—300 r.m.s.			D.C. plate current—0.4 Ma.					1247
CK501	Pentode Voltage Amplifier	—[1]	[2]	1.25	0.033	—	—	—	Class-A Amp.	30 45	0 −1.25	30 45	0.06 0.055	0.3 0.28	1000000 1500000	325 300	—	—	—	CK501
CK502	Pentode Output Amplifier	—[1]	[2]	1.25	0.033	—	—	—	Class-A Amp.	30	0	30	0.13	0.55	500000	400	—	60000	0.003	CK502
CK503	Pentode Output Amplifier	—[1]	[2]	1.25	0.033	—	—	—	Class-A Amp.	30	0	30	0.33	1.5	150000	600	—	20000	0.006	CK503
CK504	Pentode Output Amplifier	—[1]	[2]	1.25	0.033	—	—	—	Class-A Amp.	30	−1.25	30	0.09	0.4	500000	350	—	60000	0.003	CK504
CK505	Pentode Voltage Amplifier	—[1]	[2]	0.625	0.03	—	—	—	Class-A Amp.	30 45	0 −1.25	30 45	0.07 0.08	0.17 0.2	1100000 2000000	140 150	—	—	—	CK505
CK506	Pentode Output Amplifier	—[1]	[2]	1.25	0.05	—	—	—	Class-A_1 Amp.	45	−4.5	45	0.4	1.25	120000	500	—	30000	0.025	CK506
CK507	Pentode Output Amplifier	—[1]	[2]	1.25	0.05	—	—	—	Class-A_1 Amp.	45	−2.5	45	0.21	0.6	360000	500	—	50000	0.010	CK507
CK509	Triode Voltage Amplifier	—[1]	[2]	0.625	0.03	—	—	—	Class-A Amp.	45	0	—	—	0.15	150000	160	16	1000000	—	CK509
CK510	Dual Space-Charge Tetrode	—[1]	[2]	0.625	0.05	—	—	—	Class-A Amp.	45	0	0.2	200 μα	60 μα	500000	65	32.5	—	—	CK510
CK512	Low Microphonic Pentode	[1]	[2]	0.625	0.02	—	—	—	Voltage Amp.	22.5	0	22.5	0.04	0.125	—	160	—	—	—	CK412
CK515BX	Triode Voltage Amplifier	—[1]	[2]	0.625	0.03	—	—	—	Class-A Amp.	45	0	—	—	0.15	—	160	24	1000000	—	CK515BX
CK520AX	Audio Pentode	[1]	[2]	0.625	0.05	—	—	—	Class-A_1 Amp.	45	−2.5	45	0.07	0.24	—	180	—	—	0.0045	CK520AX
CK521AX	Audio Pentode	[1]	[2]	1.25	0.05	—	—	—	Class-A_1 Amp.	22.5	−3	22.5	0.22	0.8	—	400	—	—	0.006	CK521AX
CK522AX	Audio Pentode	[1]	[2]	1.25	0.02	—	—	—	Class-A_1 Amp.	22.5	0	22.5	0.08	0.3	—	450	—	—	0.0012	CK522AX
CK523AX	Pentade Output Amp.	[1]	—	1.25	0.03	—	—	—	Class-A Amp.	22.5	−1.2	22.5	0.075	0.3	—	360	—	—	0.0025	CK523AX
CK524AX	Pentode Output Amp.	[1]	—	1.25	0.03	—	—	—	Class-A Amp.	15	−1.75	15	0.125	0.45	—	300	—	—	0.0022	CK524AX
CK525AX	Pentode Output Amp.	[1]	—	1.25	0.2	—	—	—	Class-A Amp.	22.5	−1.2	22.5	0.06	0.25	—	325	—	—	0.0022	CK525AX
CK526AX	Pentode Output Amp.	[1]	—	1.25	0.2	—	—	—	Class-A Amp.	22.5	−1.5	22.5	0.12	0.45	—	400	—	—	0.004	CK526AX
CK527AX	Pentode Output Amp.	[1]	—	1.25	0.015	—	—	—	Class-A Amp.	22.5	0	22.5	0.025	0.1	—	75	—	—	0.0007	CK527AX
CK529AX	Shielded Output Pentode	[1]	—	1.25	0.02	—	—	—	Class-A Amp.	15	−1.5	15	0.05	0.2	—	275	—	—	0.0012	CK529AX
CK551AXA	Diode Pentode	[1]	[2]	1.25	0.03	—	—	—	Detector-Amp.	22.5	0	22.5	0.04	0.17	—	235	—	—	—	CK551AXA
CK553AXA	R.F. Pentode	[1]	[2]	1.25	0.05	—	—	—	Class-A_1 Amp.	22.5	0	22.5	0.13	0.42	—	550	—	—	—	CK553AXA
CK556AX	U.h.f. Triode	[1]	[2]	1.25	0.125	—	—	—	R.F. Oscillator	135	−5	—	—	4.0	—	1600	—	—	—	CK556AX
CK568AX	U.h.f. Triode	[1]	[2]	1.25	0.07	—	—	—	R.F. Oscillator	135	−6	—	—	1.9	—	650	—	—	—	CK568AX
CK569AX	R.F. Pentode	[1]	[2]	1.25	0.05	—	—	—	Class-A_1 Amp.	67.5	0	67.5	0.48	1.8	—	1100	—	—	—	CK569AX
CK605CX	Sharp Cut-off Pentode	[1]	—	6.3	0.2	—	—	—	Class-A Amp.	120	−2	120	2.5	7.5	—	5000	—	—	—	CK605CX
CK606BX	Single Diode	[1]	[2]	6.3	0.15	—	—	—	Detector	150 a.c.	—	—	—	9.0 d.c.	—	—	—	—	—	CK606BX
CK608CX	U.h.f. Triode	[1]	[2]	6.3	0.2	—	—	—	500-Mc. Osc.	120	−2	—	—	9.0	—	5000	—	—	0.75	CK608CX
CK619CX	Hi-Mu Triode	[1]	[2]	6.3	0.2	—	—	—	Class-A_1 Amp.	250	−2	—	—	4.0	—	4000	—	—	—	CK619CX
CK624CX	Sharp Cut-off Pentode	[1]	—	6.3	0.2	—	—	—	Class-A Amp.	120	−2	120	3.5	5.2	—	3000	—	—	—	CK624CX
CK650AX	Sharp Cut-off Pentode	[1]	[2]	6.3	0.2	—	—	—	Class-A_1 Amp.	120	−2	120	2.5	7.5	—	5000	—	—	—	CK650AX
CK5672	Pentode Output Amp.	[1]	—	1.25	0.05	—	—	—	Class-A Amp.	67.5	−6.25	67.5	1.0	2.75	—	625	—	—	0.06	CK5672
HY113 HY123	Triode Amplifier	—[1]	5K	1.4	0.07	—	—	—	Class-A Amp.	45	−4.5	—	—	0.4	25000	250	6.3	40000	0.0065	HY113 HY123
HY115 HY145	Pentode Voltage Amplifier	—[1]	5K	1.4	0.07	—	—	—	Class-A Amp.	45 90	−1.5 −1.5	22.5 45	0.008 0.1	0.03 0.48	5200000 1300000	58 270	300 370	—	—	HY115 HY145
HY125 HY155	Pentode Power Amplifier	—[1]	5K	1.4	0.07	—	—	—	Class-A Amp.	45 90	−3.0 −7.5	45 90	0.2 0.5	0.9 2.6	825000 420000	310 450	255 190	50000 28000	0.0115 0.09	HY125 HY155
M54	Tetrode Power Amplifier	[1]	[2]	0.625	0.04	—	—	—	Class-A Amp.	30	0	30	0.06	0.5	130000	200	26	35000	0.005	M54
M64	Tetrode Voltage Amplifier	[1]	[2]	0.625	0.02	—	—	—	Class-A Amp.	30	0	—	—	0.03	200000	110	25	—	—	M64
M74	Tetrode Voltage Amplifier	[1]	[2]	0.625	0.02	—	—	—	Class-A Amp.	30	0	7.0	0.01	0.02	500000	125	70	—	—	M74
RK61	Gas Triode	[1]	[2]	1.4	0.05	—	—	—	Radia Control	45	—	—	—	1.5	—	—	—	—	—	RK61
SD917A 5637	Triode	1	2	6.3	0.15	2.6	0.7	1.4	Class-A_1 Amp.	100	820*	—	—	1.4	26000	2700	70	—	—	SD917A 5637

TABLE XII — SUB-MINIATURE TUBES — *Continued*

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
SD828A 5638	Audio Pentode	[1]	[2]	6.3	0.15	4.0	3.0	0.22	Class-A_1 Amp.	100	270*	100	1.25	4.8	150000	3300	—	—	—	SD828A 5638
SD828E 5634	Sharp Cut-off Pentode	[4]	—	6.3	0.15	4.4	2.8	0.01	Class-A_1 Amp.	100	150*	100	2.5	6.5	240000	3500	—	—	—	SD828E 5634
SN944 5633	Remote Cut-off Pentode	[4]	—	6.3	0.15	4.0	2.8	0.01	Class-A_1 Amp.	100	150*	100	2.8	7.0	200000	3400	—	—	—	SN944 5633
SN946	Diode	[1]	[2]	6.3	0.15	1.8	—	—	Rectifier	150	—	—	—	9.0	—	—	—	—	—	SN946
SN947D 5640	Audio Beam Pentode	[1]	[2]	6.3	0.45	—	—	—	Class-A_1 Amp.	100	−9	100	2.2	31.0	15000	5000	—	3000	1.25	SN947C 5640
SN948C	Voltage Regulator	[1]	—	—	—	—	—	—	Regulator	Operating voltage = 95; Max. current = 25 Ma.										SN948C
SN953D	Power Pentode	[1]	—	6.3	0.15	9.5	3.8	0.2	Class-A Amp.	150	100*	100	4/7.5	21/20	50000	9000	—	9000	1.0	SN953D
SN954 5641	Half-Wave Rectifier	[1]	[2]	6.3	0.45	—	—	—	Rectifier	300	—	—	—	45.0	—	—	—	—	—	SN954 5641
SN955B	Dual Triode	[1]	[2]	6.3	0.45	2.8	1.0	1.3	Class-A_1 Amp. [5]	100	100*	—	—	5.5	8000	4250	34	—	—	SN955B
SN956B 5642	H.V. Half-Wave Rectifier	[4]	—	1.25	0.14	—	—	—	H.V. Rectifier	Peak inverse V. = 10000 Max. Average Ip = 2 Ma. Peak Ip = 23 Ma.										SN956B 5642
SN957A 5645	Triode	[1]	[2]	6.3	0.15	2.0	1.0	1.8	Class-A_1 Amp.	100	560*	—	—	5.0	7400	2700	20	—	—	SN957A 5645
SN1006	Triode	[1]	[2]	6.3	0.15				Class-A_1 Amp.	100	820*	—	—	1.4	29000	2400	70	—	—	SN1006
SN1007B	Mixer	[4]	—	6.3	0.15	5.0	2.8	0.003	Mixer	100	150*	100	5.0	4.0	230000	900	—	—	—	SN1007B

* Cathode resistor ohms. [1] No base; tinned wire leads. [2] Leads identified on tube. [3] No screen connection. [4] Double-ended type. [5] Values per triode.

TABLE XIII—CONTROL AND REGULATOR TUBES

Type	Name	Base	Socket Connections	Cathode	Fil. or Heater Volts	Fil. or Heater Amp.	Use	Peak Anode Voltage	Max. Anode Ma.	Minimum Supply Voltage	Operating Voltage	Operating Ma.	Grid Resistor	Tube Voltage Drop	Type
0A2	Voltage Regulator	7-pin B.	5BO	Cold	—	—	Voltage Regulator	—	—	185	150	5–30	—	—	0A2
OA5	Gas Pentode	7-pin B.	Fig. 33	Cold	—	—	Relay or Trigger	Plate —750 V., Screen —90 V., Grid +3 V., Pulse —85 V.							OA5
0B2	Voltage Regulator	7-pin B.	5BO	Cold	—	—	Voltage Regulator	—	—	133	108	5–30	—	—	0B2
0A4G 1267	Gas Triode Starter-Anode Type	6-pin O.	4V 4V	Cold	—	—	Cold-Cathode Starter-Anode Relay Tube	With 105–120-volt a.c. anode supply, peak starter-anode a.c. voltage is 70, peak r.f. voltage 55. Peak d.c. ma = 100. Average d.c. ma = 25.							0A4G 1267
1B47	Voltage Regulator	7-pin B.	—	—	—	—	Voltage Regulator	—	—	225	82	1–2	—	—	1B47
1C21	Gas Triode Glow-Discharge Type	6-pin O.	4V	Cold	—	—	Relay Tube	125–145	25	66 [6]	—	—	—	73	1C21
							Voltage Regulator		0.1 [6]	180 [4]				55	
2A4G	Gas Triode Grid Type	7-pin O.	5S	Fil.	2.5	2.5	Control Tube	200	100	—	—	—	—	15	2A4G
6Q5G	Gas Triode Grid Type	8-pin O.	6Q	Htr.	6.3	0.6	Sweep Circuit Oscillator	300	300	—	—	1.0	0.1–10 [7]	19	6Q5G
2B4		5-pin M.	5A	Htr.	2.5	1.4									2B4
2C4	Gas Triode	7-pin B.	5AS	Fil.	2.5	0.65	Control Tube	Plate volts = 350; Grid volts = —50; Avg. Ma. = 5; Peak Ma. = 20; Voltage drop = 16.							2C4
2D21	Gas Tetrode	7-pin B.	7BN	Htr.	6.3	0.6	Grid-Controlled Rectifier	650	500	—	650	100	0.1–10 [7]	8	2D21
							Relay Tube	400	—	—	—	—	1.0 [7]	—	
3C23	Gas and Mercury Vapor Grid Type	4-pin M.	3G	Fil.	2.5	7.0	Grid-Controlled Rectifier	1000	6000	—	500	1500	−4.5 [8]	15	3C23
										—	100	1500	−2.5 [8]	15	
6D4	Gas Triode	7-pin B.	5AY	Htr.	6.3	0.25	Control Tube	Plate volts = 350; Grid volts = —50; Avg. Ma. = 25; Peak Ma. = 100; Voltage drop = 16.							6D4
17	Mercury Vapor Triode	4-pin M.	3G	Fil.	2.5	5.0	Grid-Controlled Rectifier	7500 [5]	2000	—	—	500	200–3000	—	17
								2500		−5 [3]	1000	250	—	10–24	
874	Voltage Regulator	4-pin M.	4S	—	—	—	Voltage Regulator	—	—	125	90	10–50	—	—	874
876	Current Regulator	Mogul	—	—	—	—	Current Regulator	—	—	—	40–60	1.7	—	—	876
884	Gas Triode Grid Type	6-pin O.	6Q	Htr.	6.3	0.6	Sweep Circuit Oscillator	300	300	—	—	2	25000	—	884
							Grid-Controlled Rectifier	350	300	—	—	75	25000	—	
885	Gas Triode Grid Type	5-pin S.	5A	Htr.	2.5	1.4	Same as Type 884	Characteristics same as Type 884							805

TABLE XIII—CONTROL AND REGULATOR TUBES

Type	Name	Base	Socket Connec-tions	Cathode	Fil. or Heater Volts	Fil. or Heater Amp.	Use	Peak Anode Voltage	Max. Anode Ma.	Minimum Supply Voltage	Operating Voltage	Operating Ma.	Grid Resistor	Tube Voltage Drop	Type
886	Current Regulator	Mogul	—	—	—	—	Current Regulator	—	—	—	40–60	2.05	—	—	886
967	Mercury Vapor Triode	4-pin M.	3G	Fil.	2.5	5.0	Grid-Controlled Rectifier	2500	500	−5[3]	—	—	—	10–24	967
991	Voltage Regulator	Bayonet	—	—	—	—	Voltage Regulator	—	—	87	55–60	2.0	—	—	991
1265	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	130	90	5–30	—	—	1265
1266	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	—	70	5–40	—	—	1266
1267	Gas Triode	6-pin O.	4V	Cold	—	—	Relay Tube			Characteristics same as OA4G					1267
2050	Gas Tetrode	8-pin O.	8BA	Htr.	6.3	0.6	Grid-Controlled Rectifier	650	500	—	—	100	0.1–10[7]	8	2050
2051	Gas Tetrode	8-pin O	8BA	Htr.	6.3	0.6	Grid-Controlled Rectifier	350	375	—	—	75	0.1–10[7]	14	2051
2523N1/ 128AS	Gas Triode Grid Type	5-pin M.	5A	Htr.	2.5	1.75	Relay Tube	400	300	—	—	1.0	300[7]	13	2523N1/ 128AS
5651	Voltage Regulator	7-pin B.	5BO	Cold	—	—	Voltage Regulator	115	—	115	87	1.5–3.5	—	—	5651
KY21	Gas Triode Grid Type	4-pin M.	—	Fil.	2.5	10.0	Grid-Controlled Rectifier	—	—	—	3000	500	—	—	KY21
RK61	Thyratron	—[9]	—	Fil.	1.4	0.05	Radio-Controlled Relay	45	1.5	30	—	0.5–1.5	3[6]	30	RK61
RK62	Gas Triode Grid Type	4-pin S.	4D	Fil.	1.4	0.05	Relay Tube	45	1.5	—	30–45	0.1–1.5	—	15	RK62
RM208	Permatron	4-pin M.	—	Fil.	2.5	5.0	Controlled Rectifier[1]	7500[2]	1000	—	—	—	—	15	RM208
RM209	Permatron	4-pin M.	—	Fil.	5.0	10.0	Controlled Rectifier[1]	7500[2]	5000	—	—	—	—	15	RM209
OA3/VR75	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	105	75	5–40	—	—	OA3/VR75
OB3/VR90	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	125	90	5–40	—	—	OB3/VR90
OC3/VR105	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	135	105	5–40	—	—	OC3/VR105
OD3/VR150	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	185	150	5–40	—	—	OD3/VR150
KY866	Mercury Vapor Triode	4-pin M.	Fig. 8	Fil.	2.5	5.0	Grid-Controlled Rectifier	10000	1000	0–150	—	—	—	—	KY866

[1] For use as grid-controlled rectifier or with external magnetic control. RM-208 has characteristics of 866, RM-209 of 872.
[2] When under control peak inverse rating is reduced to 2500.
[3] At 1000 anode volts.
[4] Grid tied to plate.
[5] Peak inverse voltage.
[6] Grid.
[7] Megohms.
[8] Grid voltage.
[9] No base. Tinned wire leads.

TABLE XIV—CATHODE-RAY TUBES AND KINESCOPES

Type	Name	Socket Connec-tions	Heater Volts	Heater Amp.	Use	Size	Anode No. 2 Voltage	Anode No. 1 Voltage	Cut-Off Grid Voltage	Grid No. 2 Voltage	Signal-Swing Voltage	Max. Input Voltage[1]	Screen Input Power[2]	Deflection Sensitivity[6] $D_1 D_2$	Deflection Sensitivity[6] $D_3 D_4$	Anode No. 3 Voltage	Pattern Color	Type
2AP1[7]-11	Electrostatic Cathode-Ray	11B	6.3	0.6	Oscillograph Television	2"	1000	250	− 60	—	—	660	—	0.11	0.13	—	Green	2AP1-11
							500	125	− 30	—	—			0.22	0.26			
2BP1-11	Electrostatic Cathode-Ray	12E	6.3	0.6	Oscillograph	2"	2000	300/560	−135	—	—	500	—	270[3]	174[3]	—	Green	2BP1 11
							1000	150/280	−67.5	—	—	500	—	135[3]	87[3]	—		
3AP1/ 906-P1-4-5-11[7]	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	3"	1500	430	− 50	—	—	550	10	0.22	0.23		Green Blue White	3AP1/ 906-P1-4-5-11
							1000	285	− 33	—	—			0.33	0.35			
							600	170	− 20	—	—			0.55	0.58			
3BP1-4-11	Electrostatic Cathode-Ray	14A	6.3	0.6	Oscillograph	3"	2000	575	− 60	—	—	550	—	0.13	0.17	—	Green	3BP1-4-11
							1500	430	− 45	—	—			0.17	0.23			
3DP1	Electrostatic Cathode-Ray	Fig. 49	6.3	0.6	Oscillograph	3"	2000	575	− 60	—	—	550	—	200[3]	148[3]	—	Green	3DP1
							1500	430	− 40	—	—			150[3]	111[3]	—		
3EP1/ 1806-P1	Electrostatic Cathode-Ray	11A	6.3	0.6	Oscillograph Television	3"	2000	575	− 60	—	—	550	—	0.115	0.154	—	Green	3EP1/ 1806-P1
							1500	430	− 45	—	—			0.153	0.205			
3GP1-4-5-11	Electrostatic Cathode-Ray	11A	6.3	0.6	Oscillograph	3"	1500	350	− 50	—	—	550	—	0.21	0.24	—	White Green Blue	3GP1-4-5-11
							1000	234	− 33	—	—		—	0.32	0.36			
3JP1-2-4-7-11	Electrostatic Cathode-Ray	14B	6.3	0.6	Oscillograph	3"	2000	575	− 60	—	—	550	—	0.13	0.17	4000	Green Blue White	3JP1-2-4-7-11
							1500	430	− 45	—	—		—	0.17	0.23	3000		
3KP1	Electrostatic Cathode-Ray	11M	6.3	0.6	Oscillograph	3"	1000	300	− 45	1000	—	500	—	68[3]	136[3]	—	Green	3KP1
							2000	600	− 90	2000	—			52[3]	104[3]			

TABLE XIV—CATHODE-RAY TUBES AND KINESCOPES—*Continued*

Type	Name	Socket Connections	Heater Volts	Heater Amp.	Use	Size	Anode No. 2 Voltage	Anode No. 1 Voltage	Cut-Off Grid Voltage	Grid No. 2 Voltage	Signal-Swing Voltage	Max. Input Voltage [1]	Screen Input Power [2]	Deflection Sensitivity [6] $D_1 D_2$	Deflection Sensitivity [6] $D_3 D_4$	Anode No. 3 Voltage	Pattern Color	Type
3MP1	Electrostatic Cathode-Ray	Fig. 2	6.3	0.6	Oscillograph	3″	1000	200/350	−68	—	—			190[3]	180[3]	—	Green	3MP1
3RP1	Electrostatic Cathode-Ray	12E	6.3	0.6	Oscillograph	3″	1000	165/310	−67.5	—	—	—	—	73/99[3]	52/70[3]	—	Green	3RP1
							2000	330/620	−135	—	—	—	—	146/190[3]	104/140[3]	—		
5AP1/ 1805-P1 5AP4/ 1805-P4[7]	Electrostatic Picture Tube	11A	6.3	0.6	Oscillograph Television	5″	2000	575	−35	—	—	500	10	0.17	0.21	—	Green White	5AP1/ 1805-P1 5AP4/ 1805-P4
							1500	430	−27					0.23	0.28	—		
5BP1/ 1802-P1-2-4-5-11	Electrostatic Picture Tube	11A	6.3	0.6	Oscillograph	5″	2000	450	−40	—	—	500	10	0.3	0.33	—	Green White Blue	5BP1/ 1802-P1-2-4-5-7-11
							1500	337	−30	—	—			0.4	0.45			
5CP1-2-4-5-7-11	Electrostatic Cathode-Ray	14B	6.3	0.6	Oscillograph Television	5″	2000	575	−60	—	—	550	—	0.28	0.32	4000	White Green Blue	5CP1-2-4-5-11
							1500	430	−45	—	—		—	0.37	0.43	3000		
							2000	575	−60	—	—		—	0.36	0.41	2000		
5FP1-2-4-11[7]	Electromagnetic Cathode-Ray	5AN	6.3	0.6	Oscillograph Television	5″	7000	250	−45	—	—	—	—	—	—	—	Green White Blue	5FP1-2-4-11
							4000	250	−45	—	—		—	—	—			
5HP1 5HP4[7]	Electrostatic Cathode-Ray	11A	6.3	0.6	Oscillograph	5″	2000	425	−40	—	—	500	—	0.3	0.33	—	Green White	5HP1 5HP4
							1500	310	−30					0.4	0.44	—		
5JP1-2-4-5-11	Electrostatic Cathode-Ray	11E	6.3	0.6	Oscillograph	5″	2000	520	−75	—	—	500	—	0.25	0.28	4000	White Green Blue	5JP1-2-4-5-11
							1500	390	−56	—	—		—	0.33	0.37	3000		
5LP1-2-4-5-11	Electrostatic Cathode-Ray	11F	6.3	0.6	Oscillograph Television	5″	2000	500	−60	—	—	500	—	0.25	0.28	4000	White Green Blue	5LP1-2-4-5-11
							1500	375	−45	—	—		—	0.33	0.37	3000		
							1000	250	−30	—	—		—	0.49	0.56	2000		
5MP1-4-5-11	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	5″	1500	375	−50	—	—	660	—	0.39	0.42	—	White Green Blue	5MP1-4-5-11
							1000	250	−33	—	—		—	0.58	0.64			
5RP1-2-4-11	Electrostatic Cathode-Ray	14F	6.3	0.6	Oscillograph	5″	3000	—	−90	—	—	1200	—	0.12	0.12	15000	Green White Blue	5RP1-2-4-11
							2000	575	−60	—	—		—	0.18	0.18	10000		
5TP4	Projection Kinescope	12C	6.3	0.6	Television	5″	27000	4900	−70	200	—	—	—	—	—	—	White	5TP4
5UP1-7-11	Electrostatic Cathode-Ray	12E	6.3	0.6	Oscillograph	5″	2500	640	−90	—	—	500	—	38.5[3]	77[3]	—	Green Yel-low Blue	5UP1-7-11
							2500	340	−90	—	—	500	—	28[3]	56[3]	—		
							1000	320	−45	—	—	500	—	31[3]	62[3]	—		
							1000	170	−45	—	—	500	—	23[3]	46[3]	—		
5WP11	Transcriber Kinescope	12C	6.3	0.6	Television	5″	27000	5400	−42/−98	200	—	—	—	—	—	—	Blue	5WP11
5WP15	Flying-Spot Cathode-Ray	12C	6.3	0.6	Vid. Sig. Gen.	5″	20000	3000/3800	−42/−98	200	—	—	—	—	—	—	Blue Green	5WP15
7AP4	Electromagnetic Picture Tube	5AJ	2.5	2.1	Television	7″	3500	1000	−67.5	—	—	—	2.5	—	—	—	White	7AP4
7BP1-2-4-7-11	Electromagnetic Cathode-Ray	5AN	6.3	0.6	Oscillograph Television	7″	7000	250	−45	—	—	—	—	—	—	—	White Green Blue	7BP1-2-4-7-11
							4000	250	−45	—	—							
7CP1/[5] 1811-P1	Electromagnetic Cathode-Ray	6AZ	6.3	0.6	Oscillograph	7″	7000	1470	−45	250	—	—	—	—	—	—	Green	7CP1/ 1811-P1
							4000	840	−45	250	—			—	—			
7DP4	Kinescope	12C	6.3	0.6	Television	7″	6000	1430	−45	250	—	—	—	—	—	—	White	7DP4
7EP4	Electrostatic Cathode-Ray	11N	6.3	0.6	Television	7″	2500	650	−60	—	38	—		110[3]	95[3]	—	White	7EP4
7GP4[5]	Electrostatic Kinescope	Fig. 47	6.3	0.6	Television	7″	3000	1200	−84	3000	—	—	—	123[3]	102[3]	—	White	7GP4
7JP4	Electrostatic Kinescope	14G	6.3	0.6	Television	7″	6000	2400	−168					246[3]	204[3]	—	White	7JP4
9AP4/ 1804-P4	Electromagnetic Kinescope	6AL	2.5	2.1	Television	9″	7000	1425	−40	250	25	—	10	—	—	—	White	9AP4/ 1804-P4
							6000	1225	−38									
9CP4	Electromagnetic Kinescope	4AF	2.5	2.1	Television	9″	7000	—	−110	—	25	—	10	—	—	—	White	9CP4

TABLE XIV—CATHODE-RAY TUBES AND KINESCOPES—*Continued*

Type	Name	Socket Connections	Heater Volts	Heater Amps.	Use	Size	Anode No. 2 Voltage	Anode No. 1 Voltage	Cut-off Grid Voltage	Grid No. 2 Voltage	Signal-Swing Voltage	Max. Input Voltage [1]	Screen Input Power [2]	Deflection Sensitivity [6] D1 D2	Deflection Sensitivity [6] D3 D4	Anode No. 3 Voltage	Pattern Color	Type
9JP1/ 1809-P1	Electrostatic-Magnetic Cathode-Ray	8BR	2.5	2.1	Oscillograph	9″	5000	1570	− 90	—	—	3000	—	0.136	—	—	Green	9JP1/ 1809-P1
							2500	785	− 45					0.272	—			
10BP4	Magnetic Kinescope	12D	6.3	0.6	Television	10″	—	9000	− 45	250	—	—	—	—	—	—	—	10BP4
10EP4	Magnetic-Focus Cathode-Ray	12D	6.3	0.6	Television	10½″	—	8000	− 45	250	38	—	—	—	—	—	White	10EP4
10FP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	10″	—	9000	−27/−63	250	—	—	—	—	—	—	White	10FP4
10KP7	Magnetic Cathode-Ray	12D	6.3	0.6	Oscillograph	10″	—	9000	−27/ 63	250	—	—	—	—	—	—	—	10KP7
12AP4/ 1803-P4	Electromagnetic Picture Tube	6AL	2.5	2.1	Television	12″	7000	1460	− 75	250	25	—	10	—	—	—	White	12AP4/ 1803-P4
							6000	1240										
12CP4	Electromagnetic Picture Tube	4AF	2.5	2.1	Television	12″	7000	—	−110	—	25	—	10	—	—	—	White	12CP4
12DP4-7	Electromagnetic Cathode-Ray	5AN	6.3	0.6	Television	12″	7000	250	− 45	—	—	—	—	—	—	—	White	12DP4
							4000	250	− 45	—	—	—	—	—	—	—		
12KP4	Electromagnetic Cathode-Ray	12D	6.3	0.6	Television	12″	—	10000	−27/−63	250	—	—	—	—	—	—	White	12KP4
12LP4	Electromagnetic Kinescope	12D	6.3	0.6	Television	12″	—	11000	−27/−63	250	—	—	—	—	—	—	White	12LP4
15AP4	Electromagnetic Cathode-Ray	12D	6.3	0.6	Television	15″	—	8000	− 45	250	38	—	—	—	—	—	White	15AP4
16AP4	Electromagnetic Kinescope	12D	6.3	0.6	Television	16″	—	12000	−33/−77	300	—	—	—	—	—	—	White	16AP4
20BP4	Electromagnetic Cathode-Roy	12D	6.3	0.6	Television	20″	—	15000	− 45	250	38	—	—	—	—	—	White	20BP4
902 [7]	Electrostatic Cathode-Roy	Fig. 1	6.3	0.6	Oscillograph	2″	600	150	− 60	—	—	350	5	0.19	0.22	—	Green	902
903 [5]	Electromagnetic Cathode-Ray	6AL	2.5	2.1	Oscillograph	9″	7000	1360	−120	250	—	—	10	—	—	—	Green	903
904	Electrostatic-Magnetic Cathode-Ray	Fig. 3	2.5	2.1	Oscillograph	5″	4600	970	− 75	250	—	4000	10	0.09	—	—	Green	904
905 [7]	Electrostatic Cathode-Ray	Fig. 6	2.5	2.1	Oscillograph	5″	2000	450	− 35	—	—	1000	10	0.19	0.23	—	Green	905
907	Electrostatic Cathode-Ray	Fig. 6	2.5	2.1	Oscillograph	5″		Characteristics same as Type 905						—	—	—	Blue	907
908 [7]	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	3″		Characteristics same as Type 3AP1/906P1						—	—	—	Blue	908
908-A	Electrostatic Cathode-Ray	7CE	2.5	2.1	Oscillograph	3″	1500	430	− 50	—		500	—	0.223	0.233	—	Blue	908-A
							1000	287	− 33	—		500	—	0.334	0.348	—		
909 [5]	Electrostatic Cathode-Ray	Fig. 6	2.5	2.1	Oscillograph	5″		Characteristics same as Type 905						—	—	—	Blue	909
910 [5]	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	3″		Characteristics same as Type 3AP1/906P1						—	—	—	Blue	910
911 [5]	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	3″		Characteristics same as Type 3AP1/906P1						—	—	—	Green	911
912	Electrostatic Cathode-Ray	Fig. 8	2.5	2.1	Oscillograph	5″	10000	2000	− 66	250	—	7000	10	0.041	0.051	—	Green	912
913	Electrostatic Cathode-Ray	Fig. 1	6.3	0.6	Oscillograph	1″	500	100	− 65	—	—	250	5	0.07	0.10	—	Green	913
914 [7]	Electrostatic Cathode-Ray	Fig. 12	2.5	2.1	Oscillograph	9″	7000	1450	− 50	250	—	3000	10	0.073	0.093	—	Green	914
1800 [5]	Electromagnetic Kinescope	6AL	2.5	2.1	Television	9″	6000	1250	− 75	250	25	—	10	—	—	—	Yellow	1800
1801 [5]	Electromagnetic Kinescope	Fig. 13	2.5	2.1	Television	5″	3000	450	− 35	—	20	—	10	—	—	—	Yellow	1801
2001	Electrostatic Cathode-Ray	4AA	6.3	0.6	Oscillograph	1″				Characteristics essentially same as 913								2001
2002	Electrostatic Cathode-Ray	Fig. 1	6.3	0.6	Oscillograph	2″	600	120	—	—	—	—	—	0.16	0.17	—	Green	2002
2005	Electrostatic Cathode-Ray	Fig. 1 [4]	2.5	2.1	Television	5″	2000	1000	− 35	200	—	—	10	0.5	0.56	—	—	2005
24-XH	Electrostatic Cathode-Ray	Fig. 1	6.3	0.6	Oscilloscope	2″	600	120	− 60	—	—	—	10	0.14	0.16	—	Blue	24-XH

[1] Between Anode No. 2 and any deflecting plate.
[2] In mw./sq. cm., max.
[3] D.c. Volts/in.
[4] Cathode connected to Pin 7.
[5] Discontinued.
[6] In mm./volt d.c.
[7] Superseded by same type with suffix "A."

TABLE XV—RECTIFIERS—RECEIVING AND TRANSMITTING

See also Table XIII—Control and Regulator Tubes

Type No.	Name	Base	Socket Connections	Cathode	Fil. or Heater Volts	Fil. or Heater Amp.	Max. A.C. Voltage Per Plate	D.C. Output Current Ma.	Max. Inverse Peak Voltage	Peak Plate Current Ma.	Type
BA	Full-Wave Rectifier	4-pin M.	4J	Cold	—	—	350	350	Tube drop 80 v.		G
BH	Full-Wave Rectifier	4-pin M.	4J	Cold	—	—	350	125	Tube drop 90 v.		G
BR	Half-Wave Rectifier	4-pin M.	4H	Cold	—	—	300	50	Tube drop 60 v.		G
CE-220	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	3.0	—	20	20000	100	HV
OY4	Half-Wave Rectifier	5-pin O.	4BU	Cold	Connect Pins 7 and 8		95	75	300	500	G
OZ4	Full-Wave Rectifier	5-pin O.	4R	Cold	—	—	350	30–75	1250	200	G
1	Half-Wave Rectifier	4-pin S.	4G	Htr.	6.3	0.3	350	50	1000	400	MV
1-V	Half-Wave Rectifier	4-pin S.	4G	Htr.	6.3	0.3	350	50	—	—	HV
1B3GT/8016	Half-Wave Rectifier	6-pin O.	3C	Fil.	1.25	0.2	—	2.0	4000	17	HV
1B48	Half-Wave Rectifier	7-pin B.	—	Cold	—	—	800	6	2700	50	G
1X2	Half-Wave Rectifier	9-pin B.	Fig. 29	Fil.	1.25	0.2	—	1	15000	10	HV
1Z2	Half-Wave Rectifier	7-pin B.	7CB	Fil.	1.5	0.3	7800	2	20000	10	HV
2B25	Half-Wave Rectifier	7-pin B.	3T	Fil.	1.4	0.11	1000	1.5	—	9	HV
2V3G	Half-Wave Rectifier	6-pin O.	4Y	Fil.	2.5	5.0	—	2.0	16500	12	HV
2W3	Half-Wave Rectifier	5-pin O.	4X	Fil.	2.5	1.5	350	55	—	—	HV
2X2/879[10]	Half-Wave Rectifier	4-pin S.	4AB	Htr.	2.5	1.75	4500	7.5	—	—	HV
2X2-A	Half-Wave Rectifier	4-pin S.	4AB	Same as 2X2/879 but will withstand severe shock & vibration							HV
2Y2	Half-Wave Rectifier	4-pin M.	4AB	Fil.	2.5	1.75	4400	5.0	—	—	HV
2Z2/G84	Half-Wave Rectifier	4-pin M.	4B	Fil.	2.5	1.5	350	50	—	—	HV
3B24	Half-Wave Rectifier	4-pin M.	T-4A	Fil.	5.0 2.5[9]	3.0 3.0	— —	60 30	20000 20000	300 150	HV
3B25	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	—	500	4500	2000	G
3B26	Half-Wave Rectifier	8-pin O.	Fig. 31	Htr.	2.5	4.75	—	20	15000	8000	HV
DR-3B27	Half-Wave Rectifier	4-pin M.	4B	Fil.	2.5	5.0	3000	250	8500	1000	HV
5AZ4	Full-Wave Rectifier	5-pin O.	5T	Fil.	5.0	2.0	Same as Type 80				HV
5R4GY	Full-Wave Rectifier	5-pin O.	5T	Fil.	5.0	2.0	900[4] 950[7]	150[4] 175[7]	2800	650	HV
5T4	Full-Wave Rectifier	5-pin O.	5T	Fil.	5.0	3.0	450	250	1250	800	HV
5U4G	Full-Wave Rectifier	8-pin O.	5T	Fil.	5.0	3.0	Same as Type 5Z3				HV
5V4G	Full-Wave Rectifier	8-pin O.	5L	Htr.	5.0	2.0	Same as Type 83V				HV
5W4	Full-Wave Rectifier	5-pin O.	5T	Fil.	5.0	1.5	350	110	1000	—	HV
5X3	Full-Wave Rectifier	4-pin M.	4C	Fil.	5.0	2.0	1275	30	—	—	HV
5X4G	Full-Wave Rectifier	8-pin O.	5O	Fil.	5.0	3.0	Same as 5Z3				HV
5Y3G	Full-Wave Rectifier	5-pin O.	5T	Fil.	5.0	2.0	Same as Type 80				HV
5Y4G	Full-Wave Rectifier	8-pin O	5Q	Fil.	5.0	2.0	Same as Type 80				HV
5Z3	Full-Wave Rectifier	4-pin M.	4C	Fil.	5.0	3.0	500	250	1400	—	HV
5Z4	Full-Wave Rectifier	5-pin O.	5L	Htr.	5.0	2.0	400	125	1100	—	HV
6W4GT	Damper Service Half-Wave Rectifier	6-pin O.	4CG	Htr.	6.3	1.2	— 350	125 125	2000 1250	600 600	HV
6W5G	Full-Wave Rectifier	6-pin O.	6S	Htr.	6.3	0.9	350	100	1250	350	HV
6X4	Full-Wave Rectifier	7-pin B.	7CF	Htr.	6.3	0.6	325	70	1250	210	HV
6X5	Full-Wave Rectifier	6-pin O.	6S	Htr.	6.3	0.5	350	75	—	—	HV
6Y3G	Half-Wave Rectifier	5-pin O.	4AC	Htr.	6.3	0.7	5000	7.5	—	—	HV
6Y5[10]	Full-Wave Rectifier	6-pin S.	6J	Htr.	6.3	0.8	350	50	—	—	HV
6Z3	Half-Wave Rectifier	4-pin M.	4G	Fil.	6.3	0.3	350	50	—	—	HV
6Z5[10]	Full-Wave Rectifier	6-pin S.	6K	Htr.	6.3	0.6	230	60	—	—	HV
6ZY5G	Full-Wave Rectifier	6-pin O.	6S	Htr.	6.3	0.3	350	35	1000	150	HV
7Y4	Full-Wave Rectifier	8-pin L.	5AB	Htr.	6.3	0.5	350	60	—	—	HV
7Z4	Full-Wave Rectifier	8-pin L.	5AB	Htr.	6.3	0.9	450[1] 325[4]	100	1250	300	HV
12A7	Rectifier-Pentode	7-pin S.	7K	Htr.	12.6	0.3	125	30	—	—	HV
12Z3	Half-Wave Rectifier	4-pin S.	4G	Htr.	12.6	0.3	250	60	—	—	HV
12Z5	Voltage Doubler	7-pin M.	7L	Htr.	12.6	0.3	225	60	—	—	HV
14Y4	Full-Wave Rectifier	8-pin L.	5AB	Htr.	12.6	0.3	450[1] 325[4]	70	1250	210	HV
14Z3	Half-Wave Rectifier	4-pin S.	4G	Htr.	12.6	0.3	250	60	—	—	HV
25A7G[10]	Rectifier-Pentode	8-pin O.	8F	Htr.	25	0.3	125	75	—	—	HV
25W4	Half-Wave Rectifier	6-pin O.	4CG	Htr.	25	0.3	350	125	1250	600	HV
25X6GT	Voltage Doubler	7-pin O.	7Q	Htr.	25	0.15	125	60	—	—	HV
25Y4GT	Half-Wave Rectifier	6-pin O.	5AA	Htr.	25	0.15	125	75	—	—	HV
25Y5[10]	Voltage Doubler	6-pin S.	6E	Htr.	25	0.3	250	85	—	—	HV
25Z3	Half-Wave Rectifier	4-pin S.	4G	Htr.	25	0.3	250	50	—	—	HV
25Z4	Half-Wave Rectifier	6-pin O.	5AA	Htr.	25	0.3	125	125	—	—	HV
25Z5	Rectifier-Doubler	6-pin S.	6E	Htr.	25	0.3	125	100	—	500	HV
25Z6	Rectifier-Doubler	7-pin O.	7Q	Htr.	25	0.3	125	100	—	500	HV
28Z5	Full-Wave Rectifier	8-pin L.	5AB	Htr.	28	0.24	450[7] 325[4]	100	—	300	HV
32L7GT	Rectifier-Tetrode	8-pin O.	8Z	Htr.	32.5	0.3	125	60	—	—	HV
35W4	Half-Wave Rectifier	7-pin B.	5BQ	Htr.	35[2]	0.15	125	100[8]	330	600	HV
35Y4	Half-Wave Rectifier	8-pin O.	5AL	Htr.	35[2]	0.15	235	60 100[8]	700	600	HV
35Z3	Half-Wave Rectifier	8-pin L.	4Z	Htr.	35	0.15	250[5]	100	700	600	HV
35Z4GT	Half-Wave Rectifier	6-pin O.	5AA	Htr.	35	0.15	250	100	700	600	HV
35Z5G	Half-Wave Rectifier	6-pin O.	6AD	Htr.	35[2]	0.15	125	60 100[8]	—	—	HV

TABLE XV—RECTIFIERS—RECEIVING AND TRANSMITTING—*Continued*

See also Table XIII—Control and Regulator Tubes

Type No.	Name	Base	Socket Connec-tions	Cathode	Fil. or Heater Volts	Fil. or Heater Amp.	Max. A.C. Voltage Per Plate	D.C. Output Current Ma.	Max. Inverse Peak Voltage	Peak Plate Current Ma.	Type
35Z6G	Voltage Doubler	6-pin O.	7Q	Htr.	35	0.3	125	110	—	500	HV
40Z5GT	Half-Wave Rectifier	6-pin O.	6AD	Htr.	40 [2]	0.15	125	60 100 [8]	—	—	HV
45Z3	Half-Wave Rectifier	7-pin B.	5AM	Htr.	45	0.075	117	65	350	390	HV
45Z5GT	Half-Wave Rectifier	6-pin O.	6AD	Htr.	45 [2]	0.15	125	60 100 [8]	—	—	HV
50X6	Voltage Doubler	8-pin L.	7AJ	Htr.	50	0.15	117	75	700	450	HV
50Y6GT	Full-Wave Rectifier	7-pin O.	7Q	Htr.	50	0.15	125	85	—	—	HV
50Y7GT	Voltage Doubler	8-pin L.	8AN	Htr.	50 [2]	0.15	117	65	700		HV
50Z6G	Voltage Doubler	7-pin O.	7Q	Htr.	50	0.3	125	150	—	—	HV
50Z7G [10]	Voltage Doubler	8-pin O.	8AN	Htr.	50	0.15	117	65	—	—	HV
70A7GT	Rectifier-Tetrode	8-pin O.	8AB	Htr.	70	0.15	125 [5]	60	—	—	HV
70L7GT	Rectifier-Tetrode	8-pin O.	8AA	Htr.	70	0.15	117	70	—	350	HV
72	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	3.0	—	30	20000	150	HV
73	Half-Wave Rectifier	8-pin O.	4Y	Fil.	2.5	4.5	—	20	13000	3000	HV
80	Full-Wave Rectifier	4-pin M.	4C	Fil.	5.0	2.0	350 [4] 500 [7]	125 125	1400	375	HV
81	Half-Wave Rectifier	4-pin M.	4B	Fil.	7.5	1.25	700	85	—	—	HV
82	Full-Wave Rectifier	4-pin M.	4C	Fil.	2.5	3.0	500	125	1400	400	MV
83	Full-Wave Rectifier	4-pin M.	4C	Fil.	5.0	3.0	500	250	1400	800	MV
83-V	Full-Wave Rectifier	4-pin M.	4AD	Htr.	5.0	2.0	400	200	1100	—	HV
84/6Z4	Full-Wave Rectifier	5-pin S.	5D	Htr.	6.3	0.5	350	60	1000	—	HV
117L7GT/ 117M7GT	Rectifier-Tetrode	8-pin O.	8AO	Htr.	117	0.09	117	75	—	—	HV
117N7GT	Rectifier-Tetrode	8-pin O.	8AV	Htr.	117	0.09	117	75	350	450	HV
117P7GT	Rectifier-Tetrode	8-pin O.	8AV	Htr.	117	0.09	117	75	350	450	HV
117Z3	Half-Wave Rectifier	7-pin B.	4BR	Htr.	117	0.04	117	90	330	—	HV
117Z4GT	Half-Wave Rectifier	6-pin O.	5AA	Htr.	117	0.04	117	90	350	—	HV
117Z6GT	Voltage Doubler	7-pin O.	7Q	Htr.	117	0.075	235	60	700	360	HV
217-A [10]	Half-Wave Rectifier	4-pin J.	4AT	Fil.	10	3.25	—	—	3500	600	HV
217-C	Half-Wave Rectifier	4-pin J.	4AT	Fil.	10	3.25	—	—	7500	600	HV
Z225	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	—	250	10000	1000	MV
249-B	Half-Wave Rectifier	4-pin M.	Fig. 53	Fil.	2.5	7.5	3180	375	10000	1500	MV
HK253	Half-Wave Rectifier	4-pin J.	4AT	Fil.	5.0	10	—	350	10000	1500	HV
705A RK-705A	Half-Wave Rectifier	4-pin W.	T-3AA	Fil.	2.5 [9] 5.0	5.0 5.0	— —	50 100	35000 35000	375 750	HV
816	Half-Wave Rectifier	4-pin S.	4P	Fil.	2.5	2.0	2200	125	7500	500	MV
836	Half-Wave Rectifier	4-pin M.	4P	Htr.	2.5	5.0	—	—	5000	1000	HV
866A/866	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	3500	250	10000	1000	MV
866B	Half-Wave Rectifier	4-pin M.	4P	Fil.	5.0	5.0	—	—	8500	1000	MV
866 Jr.	Half-Wave Rectifier	4-pin M.	4B	Fil.	2.5	2.5	1250	250 [3]	—	—	MV
HY866 Jr.	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	2.5	1750	250 [3]	5000	—	MV
RK866	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	3500	250	10000	1000	MV
871 [10]	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	2.0	1750	250	5000	500	MV
878	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	7100	5	20000	—	HV
879	Half-Wave Rectifier	4-pin S.	4P	Fil.	2.5	1.75	2650	7.5	7500	100	HV
872A/872	Half-Wave Rectifier	4-pin J.	4AT	Fil.	5.0	7.5	—	1250	10000	5000	MV
975A	Half-Wave Rectifier	4-pin J.	4AT	Fil.	5.0	10.0	—	1500	15000	6000	MV
OZ4A/ 1003	Full-Wave Rectifier	5-pin O.	4R	Cold	—	—	—	110	880	—	G
1005/ CK1005	Full-Wave Rectifier	8-pin O.	5AQ	Fil.	6.3	0.1	—	70	450	210	G
1006/ CK1006	Full-Wave Rectifier	4-pin M.	4C	Fil.	1.75	2.25	—	200	1600	—	G
CK1007	Full-Wave Rectifier	8-pin O.	T-9G	Fil.	1.0	1.2	—	110	980	—	G
CK1009/BA	Full-Wave Rectifier	4-pin M.	—	Cold	—	—	—	350	1000	—	G
1274	Full-Wave Rectifier	6-pin O.	6S	Htr.	6.3	0.6	Same as 7Y4				HV
1275	Full-Wave Rectifier	4-pin M.	4C	Fil.	5.0	1.75	Same as 5Z3				HV
1616	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	—	130	6000	800	HV
1641/ RK60	Full-Wave Rectifier	4-pin M.	T-4AG	Fil.	5.0	3.0	—	50 250	4500 2500	— —	HV
1654	Half-Wave Rectifier	7-pin B.	2Z	Fil.	1.4	0.05	2500	1	7000	6	HV
5517	Half-Wave Rectifier	7-pin B.	5BU	Cold	—	—	1200	6	—	50	G
5825	Half-Wave Rectifier	4-pin M.	4P	Fil.	1.6	1.25	—	2	60000	40	HV
8008	Half-Wave Rectifier	4-pin [6]	Fig. 11	Fil.	5.0	7.5	—	1250	10000	5000	MV
8013A	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	—	20	40000	150	HV
8016	Half-Wave Rectifier	6-pin O.	4AC	Fil.	1.25	0.2	—	2.0	10000	7.5	HV
8020	Half-Wave Kectifier	4-pin M.	4P	Fil.	5.0 5.8	5.5 6.5	10000 12500	100 100	40000 40000	750 750	HV
RK19	Full-Wave Rectifier	4-pin M.	4AT	Htr.	7.5	2.5	1250	200 [4]	3500	600	HV
RK21	Half-Wave Rectifier	4-pin M.	4P	Htr.	2.5	4.0	1250	200 [4]	3500	600	HV
RK22	Full-Wave Rectifier	4-pin M.	T-4AG	Htr.	2.5	8.0	1250	200 [4]	3500	600	HV

1 With input choke of at least 20 henrys.
2 Tapped for pilot lamps.
3 Per pair with choke input.
4 Condenser input.
5 With 100 ohms min. resistance in series with plate; without series resistor, maximum r.m.s. plate rating is 117 volts.
6 Same as 872A/872 except for heavy-duty push-type base. Filament connected to pins 2 and 3, plate to top cap.
7 Choke input.
8 Without panel lamp.
9 Using only one-half of filament.
10 Discontinued.

TABLE XVI—TRIODE TRANSMITTING TUBES

Type	Max. Plate Dissi-pation Watts	Cathode		Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.)			Max. Freq. Mc. Full Ratings	Base	Socket Connec-tions	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
		Volts	Amp.					Grid to Fil.	Grid to Plate	Plate to Fil.											
958-A	0.6	1.25	0.1	135	7	1.0	12	0.6	2.6	0.8	500	A.	5BD	Class-C Amp.-Oscillator	135	− 20	7	1.0	0.035	—	0.6
3B7 [2]	—	1.4 2.8	0.22 0.11	180	25	—	20	1.4	2.6	2.6	125	O.	7AP	Class-C Amp. (Telegraphy)	180	0	25	—	—	—	2.8
RK24	1.5	2.0	0.12	180	20	6.0	8.0	3.5	5.5	3.0	125	S.	4D	Class-C Amp.-Oscillator	180	− 45	16.5	6.0	0.5	—	2.0
6J6 [2]	1.5	6.3	0.45	300	30	16	32	2.2	1.6	0.4	250	B.	7BF	Class-C Amp. (Telegraphy) [2]	150	− 10	30	16	0.35	—	3.5
9002	1.6	6.3	0.15	250	8	2.0	25	1.2	1.4	1.1	250	B.	7TM	Class-C Amp.-Oscillator	180	− 35	7	1.5	—	—	0.5
955	1.6	6.3	0.15	180	8	2.0	25	1.0	1.4	0.6	250	A.	5BC	Class-C Amp.-Oscillator	180	− 35	7	1.5	—	—	0.5
HY114B	1.8	1.4	0.155	180	12	3.0	13	1.0	1.3	1.0	300	O.	2T	Class-C Amp.-Oscillator	180	− 30	12	2.0	0.2	—	1.4 [3]
														Class-C Amp. (Telephony)	180	− 35	12	2.5	0.3	—	1.4 [3]
3A5 [2]	2.0	1.4 2.8	0.22 0.11	150	30	5.0	15	0.9	3.2	1.0	40	B.	7BC	Class-C Amp.-Oscillator [2]	150	− 35	30	5.0	0.2	—	2.2
6F4	2.0	6.3	0.225	150	20	8.0	17	2.0	1.9	0.6	500	A.	7BR	Class-C Amp.-Oscillator	150	− 15 550* 2000 [4]	20	7.5	0.2	—	1.8
HY24	2.0	2.0	0.13	180	20	4.5	9.3	2.7	5.4	2.3	60	S.	4D	Class-C Amp. (Telegraphy)	180	− 45	20	4.5	0.2	—	2.7
														Class-C Amp. (Telephony)	180	− 45	20	4.5	0.3	—	2.5
RK33 [1, 2]	2.5	2.0	0.12	250	20	6.0	10.5	3–2	3–2	2.5	60	S.	T-7DA	Class-C Amp.-Oscillator [2]	250	− 60	20	6.0	0.54	—	3.5
12AU7 [2]	2.75 [6]	6.3	0.3	350	12 [6]	3.5 [6]	18	1.5	1.5	0.5	54	B.	9A	Class-C Amp.-Oscillator [2]	350	−100	24	7	—	—	6.0
6N4	3.0	6.3	0.2	180	12	—	32	3.1	2.35	0.55	500	B.	7CA	Class-C Amp.-Oscillator	130	—	—	—	—	—	—
HY6J5GTX	3.5	6.3	0.3	330	20	4.0	20	4.2	3.8	5.0	60	O.	6Q	Class-C Amp.-Oscillator	330	− 30	20	2.0	0.2	—	3.5
														Class-C Amp. (Telephony)	250	− 30	20	2.5	0.3	—	2.5
2C22/7193	3.5	6.3	0.3	500	—	—	20	2.2	3.6	0.7	—	O.	4AM	Class-C Amp. (Telegraphy)	—	—	—	—	—	—	—
HY615 HY-E1148	3.5	6.3	0.175	300	20	4.0	20	1.4	1.6	1.2	300	O.	T-8AG	Class-C Amp.-Oscillator	300	− 35	20	2.0	0.4	—	4.0 [3]
														Class-C Amp. (Telephony)	300	− 35	20	3.0	0.8	—	3.5 [3]
GL-446A [1] GL-446B [1]	3.75	6.3	0.75	400	20	—	45	2.2	1.6	0.02	500	O.	Fig. 19	Class-C Amp.-Oscillator	250	—	—	—	—	—	—
GL-2C44 [1] GL-464A [1]	5.0	6.3	0.75	500	40	—	—	2.7	2.0	0.1	500	O.	Fig. 17	Class-C Amp.-Oscillator	250	—	—	—	—	—	—
6C4	5.0	6.3	0.15	350	25	8.0	18	1.8	1.6	1.3	54	B.	6BG	Class-C Amp.-Oscillator	300	− 27	25	7.0	0.35	—	5.5
1626	5.0	12.6	0.25	250	25	8.0	5.0	3.2	4.4	3.4	30	O.	6Q	Class-C Amp.-Oscillator	250	− 70	25	5.0	0.5	—	4.0
2C21/ RK33 [2]	5.0	6.3	0.6	250	40	12	—	1.6	1.6	2.0	—	S.	T-7DA	Class-C Amp.-Oscillator [2]	250	− 60	40	12	1.0	—	7
6N7 [2]	5.5 [6]	6.3	0.8	350	30 [6]	5.0 [6]	35	—	—	—	10	O.	8B	Class-C Amp. Oscillator [2, 11]	350	−100	60	10	—	—	14.5
2C40	6.5	6.3	0.75	500	25	—	36	2.1	1.3	0.05	500	O.	Fig. 19	Class-C Amp.-Oscillator	250	− 5	20	0.3	—	—	0.075
5556	7.0	4.5	1.1	350	40	10	8.5	4.0	8.3	3.0	6	M.	4D	Class-C Amp. (Telegraphy)	350	− 80	35	2	0.25	—	6
														Class-C Amp. (Telephony)	300	−100	30	2	0.3	—	4
2C43	12	6.3	0.9	500	40	—	48	2.9	1.7	0.05	1250	O.	Fig. 19	Class-C Amp.-Oscillator	470	—	38 [7]	—	—	—	9 [7]
2C26A	10	6.3	1.10	—	—	—	16.3	2.6	2.8	1.1	250	O.	4BB		—	—	—	—	—	—	—
2C34/ RK34 [2]	10	6.3	0.8	300	80	20	13	3.4	2.4	0.5	250	M.	T-7DC	Class-C Amp.-Oscillator [2]	300	− 36	80	20	1.8	—	16
205D	14	4.5	1.6	400	50	10	7.2	5.2	4.8	3.3	6	M.	4D	Class-C Amp.-Oscillator	400	−112	45	10	1.5	—	10
														Class-C Amp. (Telephony)	350	−144	35	10	1.7	—	7.1
2C25	15	7.0	1.18	450	60	15	8.0	6.0	8.9	3.0	—	M.	4D	Class-C Amp.-Oscillator	450	−100	65	15	3.2	—	19
														Class-C Amp. (Telephony)	350	−100	50	12	2.2	—	12
10Y	15	7.5	1.25	450	65	15	8	4.1	7.0	3.0	8	M.	4D	Class-C Amp.-Oscillator	450	−100	65	15	3.2	—	19
														Class-C Amp. (Telephony)	350	−100	50	12	2.2	—	12

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Connec-tions	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
843	15	2.5	2.5	450	40	7.5	7.7	4.0	4.5	4.0	6	M.	5A	Class-C Amp.-Oscillator	450	−140	30	5.0	1.0	—	7.5
														Class-C Amp. (Telephony)	350	−150	30	7.0	1.6	—	5.0
RK59[2]	15	6.3	1.0	500	90	25	25	5.0	9.0	1.0	—	M.	T-4D	Class-C Amp.-Oscillator	500	− 60	90	14	1.3	—	32
HY75A	15	6.3	2.6	450	90	25	9.6	1.8	2.6	1.0	175	O.	2T	Class-C Amp. (Telegraphy)	450	−140	90	20	5.2	—	26
														Class-C Amp. (Telephony)	400	−140	90	20	5.2	—	21
HY75	15	6.3	2.5	450	80	20	10	1.8	3.8	1.0	60	O.	2T	Class-C Amp.-Oscillator	450	− 50	80	12	—	—	21[3]
														Class-C Amp. (Telephony)	450	− 60	80	12	—	—	16[3]
1602	15	7.5	1.25	450	60	15	8.0	4.0	7.0	3.0	6	M.	4D	Class-C Amp. (Telegraphy)	450	−115	55	15	3.3	—	13
														Class-C Amp. (Telephony)	350	−135	45	15	3.5	—	8.0
														Class-B Amp. Audio[7]	425	− 50	110[8]	260[9]	2.5[8]	8000	25
841	15	7.5	1.25	450	60	20	30	4.0	7.0	3.0	6	M.	4D	Class-C Amp. (Telegraphy)	450	− 34	50	15	1.8	—	15
														Class-C Amp. (Telephony)	350	− 47	50	15	2.0	—	11
10 RK10[1]	15	7.5	1.25	450	65	15	8.0	3.0	8.0	4.0	— 60	M.	4D	Class-C Amp. (Telegraphy)	450	−100	65	15	3.2	—	19
														Class-C Amp. (Telephony)	350	−100	50	12	2.2	—	12
														Class-B Audio[7]	425	− 50	55[8]	130[9]	2.5[8]	8000	25
RK100[1]	15	6.3	0.9	150	250	100	40	23	19	3.0	—	M.	T-6B	Class-C Oscillator	110	—	80	8.0	—	—	3.5
														Class-C Amplifier	110	—	185	40	2.1	—	12
TUF-20	20	6.3	2.75	750	75	20	10	1.8	3.6	0.095	250	O.	2T	Class-C Amp.-Oscillator	750	−150	75	20	1.5/2.5	—	40
1608	20	2.5	2.5	425	95	25	20	8.5	9.0	3.0	45	M.	4D	Class-C Amp. (Telegraphy)	425	− 90	95	20	3.0	—	27
														Class-C Amp. (Telephony)	350	− 80	85	20	3.0	—	18
														Class-B Amp. Audio[7]	425	− 15	190[8]	130[9]	2.2[8]	4800	50
310	20	7.5	1.25	600	70	15	8.0	4.0	7.0	2.2	6	M.	4D	Class-C Amp. (Telegraphy)	600	−150	65	15	4.0	—	25
														Class-C Amp. (Telephony)	500	−190	55	15	4.5	—	18
703-A	20	1.2	4/4.5	350	75	12	8	0.9	1.1	0.6	1400	N.	—	Class-C Amplifier	350	−120	75	12	—	—	2/2.5
801-A/801	20	7.5	1.25	600	70	15	8.0	4.5	6.0	1.5	60	M.	4D	Class-C Amp. (Telegraphy)	600	−150	65	15	4.0	—	25
														Class-C Amp. (Telephony)	500	−190	55	15	4.5	—	18
														Class-B Amp. Audio[7]	600	− 75	130	320[9]	3.0[8]	10000	45
HY801-A	20	7.5	1.25	600	70	15	8.0	4.5	6.0	1.5	60	M.	4D	Class-C Amp. (Telegraphy)	600	−200	70	15	4.0	—	30
														Class-C Amp. (Telephony)	500	−200	60	15	4.5	—	22
T20	20	7.5	1.75	750	85	25	20	4.9	5.1	0.7	60	M.	3G	Class-C Amp. (Telegraphy)	750	− 85	85	18	3.6	—	44
														Class-C Amp. (Telephony)	750	−140	70	15	3.6	—	38
TZ20	20	7.5	1.75	750	85	30	62	5.3	5.0	0.6	60	M.	3G	Class-C Amp. (Telegraphy)	750	− 40	85	28	3.75	—	44
														Class-C Amp. (Telephony)	750	−100	70	23	4.8	—	38
														Class-B Amp. Audio[7]	800	0	40/136	160[9]	1.8[8]	12000	70
15E	20	5.5	4.2	—	—	—	25	1.4	1.15	0.3	600	N.	T-4AF	—	—	—	—	—	—	—	—
3-25A3 25T	25	6.3	3.0	2000	75	25	24	2.7	1.5	0.3	60	M.	3G	Class-C Amp.-Oscillator	2000	−130	63	18	4.0	—	100
															1500	− 95	67	13	2.2	—	75
															1000	− 70	72	9	1.3	—	47
														Class-B Amp. Audio[7]	2000	− 80	16/80	270[9]	0.7[8]	55500	110
3-25D3 3C24 24G	25	6.3	3.0	2000	75	25	23	2.0 1.7	1.6 1.5	0.2 0.3	150	S.	2D	Class-C Amp.-Oscillator	2000	−170	63	17	4.5	—	100
															1500	−110	67	15	3.1	—	75
															1000	− 80	72	15	2.6	—	47
														Class-B Audio[7]	2000	− 85	16/80	290[9]	1.1[8]	55500	110
3C28	25	6.3	3.0	2000	75	25	23	2.1	1.8	0.1	100	S.	Fig. 56	Class-C Amp. Oscillator	Characteristics same as 3C24						

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissipation Watts	Cathode		Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.)			Max. Freq. Mc. Full Ratings	Base	Socket Connections	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
		Volts	Amp.					Grid to Fil.	Grid to Plate	Plate to Fil.											
3C34	25	6.3	3.0	2000	75	25	23	2.5	1.7	0.4	60	S.	3G	Class-C Amp. Oscillator	Characteristics same as 3C24						
RK11[1]	25	6.3	3.0	750	105	35	20	7.0	7.0	0.9	60	M.	3G	Class-C Amp. (Telegraphy)	750	−120	105	21	3.2	—	55
														Class-C Amp. (Telephony)	600	−120	85	24	3.7	—	38
RK12	25	6.3	3.0	750	105	40	100	7.0	7.0	0.9	60	M.	3G	Class-C Amp. (Telegraphy)	750	−100	105	35	5.2	—	55
														Class-C Amp. (Telephony)	600	−100	85	27	3.8	—	38
HK24	25	6.3	3.0	2000	75	30	25	2.5	1.7	0.4	60	S.	3G	Class-C Amp. (Telegraphy)	2000	−140	56	18	4.0	—	90
														Class-C Amp. (Telephony)	1500	−145	50	25	5.5	—	60
HY25	25	7.5	2.25	800	75	25	55	4.2	4.6	1.0	60	M.	3G	Class-C Amp. (Telegraphy)	750	− 45	75	15	2.0	—	42
														Class-C Amp. (Telephony)	700	− 45	75	17	5.0	—	39
8025	30	6.3	1.92	1000	65	—	18	2.7	2.8	0.35	500	M.	4AQ	Class-C Amp. (Grid. Mod.)	1000	−135	50	4	3.5	—	20
	20				65	20								Class-C Amp. (Telephony)	800	−105	40	10.5	1.4	—	22
	30				80	20								Class-C Amp. (Telegraphy)	1000	− 90	50	14	1.6	—	35
HY30Z[1]	30	6.3	2.25	850	90	25	87	6.0	4.9	1.0	60	M.	4BO	Class-C Amp.-Oscillator	850	− 75	90	25	2.5	—	58
														Class-C Amp. (Telephony)	700	− 75	90	25	3.5	—	47
HY31Z[2] HY1231Z[2]	30	6.3 12.6	3.5 1.7	500	150	30	45	5.0	5.5	1.9	60	M.	T-4D	Class-C Amp. (Telegraphy)	500	− 45	150	25	2.5	—	56
														Class-C Amp. (Telephony)	400	−100	150	30	3.5	—	45
316A	30	2.0	3.65	450	80	12	6.5	1.2	1.6	0.8	500	N.	—	Class-C Amp. (Telegraphy)	450	—	80	12	—	—	7.5
														Class-C Amp. (Telephony)	400	—	80	12	—	—	6.5
809	30	6.3	2.5	1000	125	—	50	5.7	6.7	0.9	60	M.	3G	Class-C Amp. (Telegraphy)	1000	− 75	100	25	3.8	—	75
														Class-C Amp. (Telephony)	750	− 60	100	32	4.3	—	55
														Class-B Amp. Audio [7]	1000	− 9	40/200	155 [9]	2.7 [8]	11600	145
1623	30	6.3	2.5	1000	100	25	20	5.7	6.7	0.9	60	M.	3G	Class-C Amp.-Oscillator	1000	− 90	100	20	3.1	—	75
														Class-C Amp. (Telephony)	750	−125	100	20	4.0	—	55
														Class-B Amp. Audio [7]	1000	− 40	30/200	230 [9]	4.2 [8]	12000	145
53A	35	5.0	12.5	15000	—	—	35	3.6	1.9	0.4	—	N.	T-4B	Oscillator at 300 Mc.	Approximately 50 watts output						
RK30[1]	35	7.5	3.25	1250	80	25	15	2.75	2.5	2.75	60	M.	2D	Class-C Amp. (Telegraphy)	1250	−180	90	18	5.2	—	85
														Class-C Amp. (Telephony)	1000	−200	80	15	4.5	—	60
800	35	7.5	3.25	1250	80	25	15	2.75	2.5	2.75	60	M.	2D	Class-C Amp. (Telegraphy)	1250	−175	70	15	4.0	—	65
														Class-C Amp. (Telephony)	1000	−200	70	15	4.0	—	50
														Class-B Amp. Audio [7]	1250	− 70	30/130	300 [9]	3.4 [8]	21000	106
1628[1]	40	3.5	3.25	1000	60	15	23	2.0	2.0	0.4	500	N.	T-4BB	Class-C Amp.-Oscillator	1000	− 65	50	15	1.7	—	35
														Class-C Amp. (Telephony)	800	−100	40	11	1.6	—	22
														Grid-Modulated Amp.	1000	−120	50	3.5	5.0	—	20
8012 GL-8012-A	40	6.3	2.0	1000	80	20	18	2.7 2.7	2.8 2.5	0.35 0.4	500	N.	T-4BB	Class-C Amp.-Oscillator	1000	− 90	50	14	1.6	—	35
														Class-C Amp. (Telephony)	800	−105	40	10.5	1.4	—	22
														Grid-Modulated Amp.	1000	−135	50	4.0	3.5	—	20
RK18[1]	40	7.5	3.0	1250	100	40	18	6.0	4.8	1.8	60	M.	3G	Class-C Amp. (Telegraphy)	1250	−160	100	12	2.8	—	95
														Class-C Amp. (Telephony)	1000	−160	80	13	3.1	—	64
RK31	40	7.5	3.0	1250	100	35	170	7.0	1.0	2.0	30	M.	3G	Class-C Amp. (Telegraphy)	1250	− 80	100	30	3.0	—	90
														Class-C Amp. (Telephony)	1000	− 80	100	28	3.5	—	70
HY40[1]	40	7.5	2.25	1000	125	25	25	6.1	5.6	1.0	60	M.	3G	Class-C Amp. (Telegraphy)	1000	− 90	125	20	5.0	—	94
														Class-C Amp. (Telephony)	850	− 90	125	25	5.0	—	82
														Grid-Modulated Amp.	1000	—	125	—	—	—	20

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Connec-tions	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
HY40Z[1]	40	7.5	2.6	1000	125	30	80	6.2	6.3	0.8	60	M.	3G	Class-C Amp. (Telegraphy)	1000	− 27	125	25	5.0	—	94
														Class-C Amp. (Telephony)	850	− 30	100	30	7.0	—	82
														Grid-Modulated Amp.	1000	—	60	—	—	—	20
T40	40	7.5	2.5	1500	150	40	25	4.5	4.8	0.8	60	M.	3G	Class-C Amp.-Oscillator	1500	−140	150	28	9.0	—	158
														Class-C Amp. (Telephony)	1250	−115	115	20	5.25	—	104
TZ40	40	7.5	2.5	1500	150	45	62	4.8	5.0	0.8	60	M.	3G	Class-C Amp.-Oscillator	1500	− 90	150	38	10	—	165
														Class-C Amp. (Telephony)	1250	−100	125	30	7.5	—	116
														Class-B Amp. Audio[7]	1500	− 9	250[8]	285[9]	6.0[8]	12000	250
HY57	40	6.3	2.25	850	110	25	50	4.9	5.1	1.7	60	M.	3G	Class-C Amp. (Telegraphy)	850	− 48	110	15	2.5	—	70
														Class-C Amp. (Telephony)	700	− 45	90	17	5.0	—	47
														Grid-Modulated Amp.	850	—	70	—	—	—	20
756[1]	40	7.5	2.0	850	110	25	8.0	3.0	7.0	2.7	—	M.	4D	Class-C Amplifier	850	—	110	25	—	—	—
830[1]	40	10	2.15	750	110	18	8.0	4.9	9.9	2.2	15	M.	4D	Class-C Amplifier	750	−180	110	18	7.0	—	55
														Grid-Modulated Amp.	1000	−200	50	2.0	3.0	—	15
3-50A4 35T	50	5.0	4.0	2000	150	50	39	4.1	1.8	0.3	100	M.	3G	Class-C Amp. (Telegraphy)	2000	−135	125	45	13	—	200
3-50D4 35TG								2.5	1.8	0.4	100	M.	2D	Class-C Amp. (Telephony)	1500	−120	100	30	5.0	—	120
														Class-B Amp. Audio[7]	2000	− 40	34/167	255[9]	4.0[8]	27500	235
8010-R	50	6.3	2.4	1350	150	20	30	2.3	1.5	0.07	350	N.	—	Class-C Amplifier	—	—	—	—	—	—	—
RK32[1]	50	7.5	3.25	1250	100	25	11	2.5	3.4	0.7	100	M.	2D	Class-C Amp. (Telegraphy)	1250	−225	100	14	4.8	—	90
														Class-C Amp. (Telephony)	1000	−310	100	21	8.7	—	70
RK35[1]	50	7.5	4.0	1500	125	20	9.0	3.5	2.7	0.4	60	M.	2D	Class-C Amp. (Telegraphy)	1500	−250	115	15	5.0	—	120
														Class-C Amp. (Telephony)	1250	−250	100	14	4.6	—	93
														Grid-Modulated Amp.	1500	−180	37	—	2.0	—	25
RK37	50	7.5	4.0	1500	125	35	28	3.5	3.2	0.2	60	M.	2D	Class-C Amp. (Telegraphy)	1500	−130	115	30	7.0	—	122
														Class-C Amp. (Telephony)	1250	−150	100	23	5.6	—	90
														Grid-Modulated Amp.	1500	− 50	50	—	2.4	—	26
3-50G2 UH50	50	7.5	3.25	1250	125	25	10.6	2.2	2.6	0.3	60	M.	2D	Class-C Amp. (Telegraphy)	1250	−225	125	20	7.5	—	115
														Class-C Amp. (Telephony)	1250	−325	125	20	10	—	115
														Grid-Modulated Amp.	1250	−200	60	2.0	3.0	—	25
UH51[1]	50	5.0	6.5	2000	175	25	10.6	2.2	2.3	0.3	60	M.	2D	Class-C Amp. (Telegraphy)	2000	−500	150	20	15	—	225
														Class-C Amp. (Telephony)	1500	−400	165	20	15	—	200
														Grid-Modulated Amp.	1500	−400	85	2.0	8.0	—	65
HK54	50	5.0	5.0	3000	150	30	27	1.9	1.9	0.2	100	M.	2D	Class-C Amp. (Telegraphy)	3000	−290	100	25	10	—	250
														Class-C Amp. (Telephony)	2500	−250	100	20	8.0	—	210
														Class-B Amp. Audio[7]	2500	− 85	20/150	360[9]	5.0	40000	275
HK154[1]	50	5.0	6.5	1500	175	30	6.7	4.3	5.9	1.1	60	M.	2D	Class-C Amp. (Telegraphy)	1500	−590	167	20	15	—	200
														Class-C Amp. (Telephony)	1250	−460	170	20	12	—	162
														Grid-Modulated Amp.	1500	−450	52	—	5.0	—	28
HK158	50	12.6	2.5	2000	200	40	25	4.7	4.6	1.0	60	M.	2D	Class-C Amp.-Oscillator	2000	−150	125	25	6.0	—	200
														Class-C Amp. (Telephony)	2000	−140	105	25	5.0	—	170
WE304A[1] 304B	50	7.5	3.25	1250	100	25	11	2.0	2.5	0.7	100	M.	2D	Class-C Amp. (Telegraphy)	1250	−200	100	—	—	—	85
														Class-C Amp. (Telephony)	1000	−180	100	—	—	—	65

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Connec-tions	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
356A	50	5.0	5.0	1500	120	35	50	2.25	2.75	1.0	60	N.	T-4BD	Class-C Amp. (Telegraphy)	1500	−60	100	—	—	—	100
														Class-C Amp. (Telephony)	1250	−100	100	35	—	—	85
808	50	7.5	4.0	1500	150	35	47	5.3	2.8	0.15	30	M.	2D	Class-C Amp. (Telegraphy)	1500	−200	125	30	9.5	—	140
														Class-C Amp. (Telephony)	1250	−225	100	32	10.5	—	105
														Class-B Amp. Audio [7]	1500	−25	30/190	220 [9]	4.8 [8]	18300	185
834	50	7.5	3.1	1250	100	20	10.5	2.2	2.6	0.6	100	M.	2D	Class-C Amp. (Telegraphy)	1250	−225	90	15	4.5	—	75
														Class-C Amp. (Telephony)	1000	−310	90	17.5	6.5	—	58
841A [1]	50	10	2.0	1250	150	30	14.6	3.5	9.0	2.5	—	M.	3G	Class-C Amplifier	—	—	—	—	—	—	85
841SW	50	10	2.0	1000	150	30	14.6	—	9.0	—	—	M.	3G	Class-C Amplifier	—	—	—	—	—	—	—
T55	55	7.5	3.0	1500	150	40	20	5.0	3.9	1.2	60	M.	3G	Class-C Amp. (Telegraphy)	1500	−170	150	18	6.0	—	170
														Class-C Amp. (Telephony)	1500	−195	125	15	5.0	—	145
811	55	6.3	4.0	1500	150	50	160	5.5	5.5	0.6	60	M.	3G	Class-C Amp. (Telegraphy)	1500	−113	150	35	8.0	—	170
														Class-C Amp. (Telephony)	1250	−125	125	50	11	—	120
														Class-B Amp. Audio [7]	1500	−9	20/200	150 [9]	3.0 [8]	17600	220
812	55	6.3	4.0	1500	150	35	29	5.3	5.3	0.8	60	M.	3G	Class-C Amp. (Telegraphy)	1500	−175	150	25	6.5	—	170
														Class-C Amp. (Telephony)	1250	−125	125	25	6.0	—	120
														Class-B Amp. Audio [7]	1500	−45	50/200	232 [9]	4.7 [8]	18000	225
RK51	60	7.5	3.75	1500	150	40	20	6.0	6.0	2.5	60	M.	3G	Class-C Amp. (Telegraphy)	1500	−250	150	31	10	—	170
														Class-C Amp. (Telephony)	1250	−200	105	17	4.5	—	96
														Grid-Modulated Amp.	1500	−130	60	0.4	2.3	—	128
RK52	60	7.5	3.75	1500	130	50	170	6.6	12	2.2	60	M.	3G	Class-C Amp. (Telegraphy)	1500	−120	130	40	7.0	—	135
														Class-C Amp. (Telephony)	1250	−120	115	47	8.5	—	102
														Class-B Amp. Audio [7]	1250	0	40/300	180 [9]	7.5 [8]	10000	250
T-60	60	10	2.5	1600	150	50	20	5.5	5.2	2.5	60	M.	2D	Class-C Amp.-Oscillator	1500	−150	150	50	9.0	—	100
826	55	7.5	4.0	1000	125	40	31	3.7	2.9	1.4	250	N.	7BO	Class-C Amp.-Oscillator	1000	−70	130	35	5.8	—	90
														Class-C Amp. (Telephony)	1000	−160	95	40	11.5	—	70
														Grid-Modulated Amp.	1000	−125	65	9.5	8.2	—	25
830B 930B	60	10	2.0	1000	150	30	25	5.0	11	1.8	15	M.	3G	Class-C Amp.-Oscillator	1000	−110	140	30	7.0	—	90
														Class-C Amp. (Telephony)	800	−150	95	20	5.0	—	50
														Class-B Amp. Audio [7]	1000	−35	20/280	270 [9]	6.0 [8]	7600	175
811-A	65	6.3	4.0	1500	175	50	160	5.9	5.6	0.7	60	M.	3G	Class-C Amp. (Telegraphy)	1500	−70	173	40	7.1	—	200
														Class-C Amp. (Telephony)	1250	−120	140	45	10.0	—	135
														Class-B Amp. Audio [7]	1500	−4.5	32/313	170 [9]	4.4 [8]	12400	340
812-A	65	6.3	4.0	1500	175	35	29	5.4	5.5	0.77	60	M.	3G	Class-C Amp. (Telegraphy)	1500	−120	173	30	6.5	—	190
														Class-C Amp. (Telephony)	1250	−115	140	35	7.6	—	130
														Class-B Audio [7]	1500	−48	28/310	270 [9]	5.0	13200	340
HY51A [1] HY51B [1]	65	7.5 10	3.5 2.25	1000	175	25	25	6.5	7.0	1.1	60	M.	3G	Class-C Amp. (Telegraphy)	1000	−75	175	20	7.5	—	131
														Class-C Amp. (Telephony)	1000	−67.5	130	15	7.5	—	104
														Grid-Modulated Amp.	1000	—	100	—	—	—	33
HY51Z [1]	65	7.5	3.5	1000	175	35	85	7.9	7.2	0.9	60	M.	4BO	Class-C Amp. (Telegraphy)	1000	−22.5	175	35	10	—	131
														Class-C Amp. (Telephony)	1000	−30	150	35	10	—	104
														Grid-Modulated Amp.	1000	—	100	—	—	—	33
5514	65	7.5	3.0	1500	175	60	145	7.8	7.9	1.0	60	M.	4BO	Class-C Amp. (Telegraphy)	1500	−106	175	60	12	—	200
														Class-C Amp. (Telephony)	1250	−84	142	60	10	—	135
														Class-B Audio [7]	1500	−4.5	350 [8]	88 [8]	6.5 [8]	10500	400

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Connec-tions	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
UH35[1]	70	5.0	4.0	1500	150	35	30	1.4	1.6	0.2	60	M.	3G	Class-C Amp. (Telegraphy)	1500	−170	150	30	7.0	—	170
														Class-C Amp. (Telephony)	1500	−120	100	30	5.0	—	120
V70 V70B	70	10	2.5	1500	140	25	14	5.0	9.0	2.3	—	J. M.	3N 3G	Class-C Amp. (Telegraphy)	1500	−215	130	6.0	3.0	—	140
														Class-C Amp. (Telephony)	1250	−250	130	6.0	3.0	—	120
V70A V70C	70	10	2.5	1500	140	20	25	5.0	9.5	2.0	—	J. M.	3N 3G	Class-C Amp. (Telegraphy)	1000	−110	140	30	7.0	—	90
														Class-C Amp. (Telephony)	800	−150	95	20	5.0	—	50
50T[1]	75	5.0	6.0	3000	100	30	12	2.0	2.0	0.4	—	M.	2D	Class-C Amplifier	3000	−600	100	25	—	—	250
3-75A3 75TH	75	5.0	6.25	3000	225	40	20	2.7	2.3	0.3	40	M.	2D	Class-C Amp. (Telegraphy)	2000	−200	150	32	10	—	225
														Class-B Amp. Audio[7]	2000	−90	50/225	350[9]	3[8]	19300	300
3-75A2 75TL						35	12	2.6	2.4	0.4			2D	Class-C Amp. (Telegraphy)	2000	−300	150	21	8	—	225
														Class-B Amp. Audio[7]	2000	−160	50/250	535[9]	5[8]	18000	350
HF-60	75	10	2.5	1600	160	—	28	5.4	5.2	1.5	30	M.	2D	Class-C Amp. (Telegraphy)	1600	−190	158	12	3.5	—	200
														Class-C Amp. (Telephony)	1250	−190	113	8	2.5	—	110
														Class-B Amp. Audio[7]	1600	−75	50/248	310[9]	3.0	13800	262
ZB-60	75	10	2.5	1600	160	40	80	6.1	5.8	1.85	30	M.	2D	Class-C Amp. (Telegraphy)	1500	−95	158	31	6.0	—	190
														Class-B Amp. Audio[7]	1500	−9	30/305	208[9]	12.5	11200	320
111H	75	10	2.5	1500	160	30	23	5.0	4.6	2.9	30	M.	2D	Class-C Amp. (Telegraphy)	1500	−200	150	18	6.0	—	170
														Class-C Amp. (Telephony)	1250	−250	110	21	8.0	—	105
														Class-B Amp. Audio[7]	1750	−62	40/270	324[9]	9.0	16000	350
HF75	75	10	3.25	2000	120	—	12.5	—	2.0	—	75	M.	2D	Class-C Oscillator-Amp.	2000	—	120	—	—	—	150
TW75	75	7.5	4.15	2000	175	60	20	3.35	1.5	0.7	60	M.	2D	Class-C Amp.-Oscillator	2000	−175	150	37	12.7	—	225
														Class-C Amp. (Telephony)	2000	−260	125	32	13.2	—	198
T-100 HF100	75	10	2.5	1500	150	30	23	4.0	4.5	2.6	30	M.	2D	Class-C Amp. (Telegraphy)	1500	−200	150	18	6.0	—	170
														Class-C Amp. (Telephony)	1250	−250	110	21	8.0	—	105
														Grid-Modulated Amp.	1500	−280	72	1.5	6.0	—	42
														Class-B Amp. Audio[7]	1750	−62	40/270	324[9]	9.0[8]	16000	350
UE-100	75	10	2.5	1750	150	30	23	3.5	4.5	1.4	30	M.	2D	Class-C Amp. (Telegraphy)	1500	−200	150	18	6.0	—	170
														Class-C Amp. (Telephony)	1250	−250	120	21	8.0	—	105
														Class-B Audio[7]	1750	−62	540[8]	—	9.0	16000	350
ZB120	75	10	2.0	1250	160	40	90	5.3	5.2	3.2	30	J.	4E	Class-C Amp. (Telegraphy)	1250	−135	160	23	5.5	—	145
														Class-C Amp. (Telephony)	1000	−150	120	21	5.0	—	95
														Grid-Modulated Amp.	1250	—	95	8.0	1.5	—	45
														Class-B Amp. Audio[7]	1500	−9	60/296	196[9]	5.0[8]	11200	300
327B	75	10.5	10.6	—	—	—	30	3.4	2.45	0.3	—	N.	T-4AD	—	—	—	—	—	—	—	—
242A	85	10	3.25	1250	150	50	12.5	6.5	13	4.0	6	J.	4E	Class-C Amp. (Telegraphy)	1250	−175	150	—	—	—	130
														Class-C Amp. (Telephony)	1000	−160	150	50	—	—	100
284D	85	10	3.25	1250	150	100	4.8	6.0	8.3	5.6	—	J.	4E	Class-C Amp. (Telegraphy)	1250	−500	150	—	—	—	125
														Class-C Amp. (Telephony)	1000	−450	150	50	—	—	100
														Class-B Amp. Audio[7]	1250	−250	30/200	—	—	11200	140
812-H	85	6.3	4.0	1750	200	45	—	5.3	5.3	0.8	30	M.	3G	Class-C Amp. (Telegraphy)	1750	−175	170	26	6.5	—	225
															1250	−125	125	25	5.0	—	116
														Class-C Amp. (Telephony)	1500	−125	165	21	6.0	—	180
															1250	−125	125	25	6.0	—	120
														Class-B Amp. Audio[7]	1500	−46	42/200	—	—	18000	225

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amp.	Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Connections	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
8005	85	10	3.25	1500	200	45	20	6.4	5.0	1.0	60	M.	3G	Class-C Amp.-Telegraphy	1500	−130	200	32	7.5	—	220
														Class-C Amp. (Telephony)	1250	−195	190	28	9.0	—	170
														Class-B Amp. Audio [7]	1500	− 70	40/310	310 [9]	4.0 [8]	10000	300
V-70-D	85	7.5	3.25	1750	200	45	—	4.5	4.5	1.7	30	M.	3G	Class-C Amp. (Telegraphy)	1750	−100	170	19	3.9	—	225
															1500	− 90	165	19	3.9	—	195
														Class-C Amp. (Telephony)	1500	− 90	165	19	3.7	—	185
															1250	− 72	127	16	2.6	—	122
RK36[1]	100	5.0	8.0	3000	165	35	14	4.5	5.0	1.0	60	M.	2D	Class-C Amp. (Telegraphy)	2000	−360	150	30	15	—	200
														Class-C Amp. (Telephony)	2000	−360	150	30	15	—	200
														Grid-Modulated Amp.	2000	−270	72	1.0	3.5	—	42
RK38[1]	100	5.0	8.0	3000	165	40	—	4.6	4.3	0.9	60	M.	2D	Class-C Amp. (Telegraphy)	2000	−200	160	30	10	—	225
														Class-C Amp. (Telephony)	2000	−200	160	30	10	—	225
														Grid-Modulated Amp.	2000	−150	80	2.0	5.5	—	60
3-100A4 100TH	100	5.0	6.3	3000	225	60	40	2.9	2.0	0.4	40	M.	2D	Class-C Amp. (Telegraphy) Class-C Amp. (Telephony)	3000	−200	165	51	18	—	400
														Grid-Modulated Amp.	3000	−400	70	3.0	7.0	—	100
														Class-B Amp. (Audio) [7]	3000	− 65	40/215	335 [9]	5.0 [8]	31000	650
3-100A2 100TL	100	5.0	6.3	3000	225	50	14	2.3	2.0	0.4	40	M.	2D	Class-C Amp. (Telegraphy) Class-C Amp. (Telephony)	3000	−400	165	30	20	—	400
														Grid-Modulated Amp.	3000	−560	60	2.0	7.0	—	90
														Class-B Amp. (Audio) [7]	3000	−185	40/215	640 [9]	6.0 [8]	30000	450
VT127A	100	5.0	10.4	3000	—	30	15.5	2.7	2.3	0.35	150	N.	T-4B	Class-C Amp.-Oscillator		Characteristics similar to 100TL					
227A	100	10.5	10.7	—	—	—	31	3.0	2.2	0.30	—	N.	T-4B	Oscillator at 200 Mc.	—	—	—	—	—	—	—
327A	100	10.5	10.7	—	—	—	31	3.4	2.3	0.35	—	N.	T-4AD	Oscillator at 200 Mc.	—	—	—	—	—	—	—
HK254	100	5.0	7.5	4000	200	40	25	3.3	3.4	1.1	50	J.	2N	Class-C Amp. (Telegraphy)	4000	−380	120	35	20	—	475
														Class-C Amp. (Telephony)	3000	−290	135	40	23	—	320
														Grid-Modulated Amp.	3000	—	51	3.0	4.0	—	58
														Class-B Amp. (Audio) [7]	3000	−100	40/240	456 [9]	7.0 [8]	30000	520
RK58	100	10	3.25	1250	175	70	—	8.5	6.5	10.5	—	J.	3N	Class-C Amp. (Telegraphy)	1250	− 90	150	30	6.0	—	130
														Class-C Amp. (Telephony)	1000	−135	150	50	16	—	100
HF120	100	10	3.25	1250	175	50	12	5.5	12.5	3.5	15	J.	4F	Class-C Amp.-Oscillator	1250	−300	166	8	3.5	—	148
HF125	100	10	3.25	1500	175	—	25	—	11.5	—	30	J.	—	Class-C Amp.-Oscillator	1500	—	175	—	—	—	200
HF140	100	10	3.25	1250	175	—	12	5.5	13.0	4.5	15	J.	4F	Class-C Amp.-Oscillator	1250	−300	166	8	3.5	—	148
203A 303A	100	10	3.25	1250	175	60	25	6.5	14.5	5.5	15	J.	4E	Class-C Amp. (Telegraphy)	1250	−125	150	25	7.0	—	130
														Class-C Amp. (Telephony)	1000	−135	150	50	14	—	100
														Class-B Amp. (Audio) [7]	1250	− 45	26/320	330 [9]	11 [8]	9000	260
203H	100	10	3.25	1500	175	60	25	6.5	11.5	1.5	15	J.	3N	Class-C Amp. (Telegraphy)	1500	−200	170	12	3.8	—	200
														Class-C Amp. (Telephony)	1250	−160	167	19	5.0	—	160
														Class-B Amp. (Audio) [7]	1500	− 52	30/320	304 [9]	5.5 [8]	11000	340
211 311 835[1]	100	10	3.25	1250	175	50	12	6.0 6.0	14.5 9.25	5.5 5.0	15	J.	4E	Class-C Amp. (Telegraphy)	1250	−225	150	18	7.0	—	130
														Class-C Amp. (Telephony)	1000	−260	150	35	14	—	100
														Class-B Amp. (Audio) [7]	1250	−100	20/320	410 [9]	8.0 [8]	9000	260
242B 342B	100	10	3.25	1250	150	50	12.5	7.0	13.6	6.0	6	J.	4E	Class-C Amp. (Telegraphy)	1250	−175	150	—	—	—	130
														Class-C Amp. (Telephony)	1000	−160	150	50	—	—	100

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amp.	Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Connections	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
242C	100	10	3.25	1250	150	50	12.5	6.1	13.0	4.7	6	J.	4E	Class-C Amp. (Telegraphy)	1250	−175	150	—	—	—	130
														Class-C Amp. (Telephony)	1000	−160	150	50	—	—	100
														Class-B Amp. (Audio)[7]	1250	−80	25/150	—	25[8]	7600	200
261A 361A	100	10	3.25	1250	150	50	12	6.5	9.0	4.0	30	J.	4E	Class-C Amp. (Telegraphy)	1250	−175	125	—	—	—	100
														Class-C Amp. (Telephony)	1000	−160	150	50	—	—	100
														Class-B Amp. (Audio)[7]	1250	−90	20/150	—	25[8]	7200	200
276A 376A	100	10	3.0	1250	125	50	12	6.0	9.0	4.0	30	J.	4E	Class-C Amp. (Telegraphy)	1250	−175	125	—	—	—	100
														Class-C Amp. (Telephony)	1000	−160	125	50	—	—	85
														Class-B Amp. (Audio)[7]	1250	−90	20/125	—	25[8]	9000	175
284B	100	10	3.25	1250	150	100	5.0	4.2	7.4	5.3	—	J.	3N	Class-C Amp. (Telegraphy)	1250	−500	150	—	—	—	125
														Class-C Amp. (Telephony)	1000	−430	150	50	—	—	100
														Class-B Amp. (Audio)[7]	1250	−245	15/150	—	10[8]	7200	200
295A	100	10	3.25	1250	175	50	25	6.5	14.5	5.5	—	J.	4E	Class-C Amp. (Telegraphy)	1250	−125	150	—	—	—	125
														Class-C Amp. (Telephony)	1000	−125	150	50	—	—	100
														Class-B Amp. (Audio)[7]	1250	−40	12/160	—	20[8]	9000	250
838 938	100	10	3.25	1250	175	70	—	6.5	8.0	5.0	30	J.	4E	Class-C Amp. (Telegraphy)	1250	−90	150	30	6.0	—	130
														Class-C Amp. (Telephony)	1000	−135	150	60	16	—	100
														Class-B Amp. (Audio)[7]	1250	0	148/320	200[9]	7.5[8]	9000	260
852	100	10	3.25	3000	150	40	12	1.9	2.6	1.0	30	M.	2D	Class-C Amp. (Telegraphy)	3000	−600	85	15	12	—	165
														Class-C Amp. (Telephony)	2000	−500	67	30	23	—	75
														Class-B Amp. (Audio)[7]	3000	−250	14/160	780[9]	3.5[8]	10250	320
5648[12]	100	6.3	1.1	1000	100	50	100	8.75	1.95	0.035	2500	N.	—	Class-C Amp. (Telegraphy)	1000	−50	50	18	4	—	30
														Class-C Amp. (Telephony)	600	−25	55	22	6	—	20
8003	100	10	3.25	1500	250	50	12	5.8	11.7	3.4	30	J.	3N	Class-C Amp.-Oscillator	1350	−180	245	35	11	—	250
														Class-C Amp. (Telephony)	1100	−260	200	40	15	—	167
														Class-B Amp. (Audio)[7]	1350	−100	40/490	480[9]	10.5[8]	6000	460
3X100A11 2C39	100	6.3	1.1	1000	60	40	100	6.5	1.95	0.03	500	N.	—	"Grid Isolation" Circuit	600	−35	60	40	5.0	—	20
311-CH	125	10	3.25	1750	200	50	12	5.5	8.0	4.5	30	J.	Fig. 57	Class-C Amp. (Telegraphy)	1750	−200	200	20	4.5	—	260
														Class-C Amp. (Telephony)	1250	−200	166	8	3.5	—	148
														Class-B (Audio)[7]	1500	−110	400[8]	—	—	8200	400
3C22	125	6.3	2.0	1000	150	70	40	4.9	2.4	0.05	500	O.	Fig. 30	Class-C Amp.-Oscillator	1000	−200	150	70	—	—	65
4C36	125	5	7 5	4000	—	—	29	3.2	3.0	0.4	60	J.	Fig. 56	Class-C Amp.-Oscillator	—	—	—	—	18	—	480
F-123-A DR-123C	125	10	4.0	2000	300	75	14.5	6.5	8.5	3.3	—	J.	Fig. 26	Class-C Amp. (Telegraphy)	1500	−250	250	30	11	—	300
														Class-C Amp. (Telephony)	1500	−290	160	25	10	—	200
														Class-B Amp. (Audio)[7]	2000	−130	30/175	217[9]	3.4[8]	13800	522
RK57/805	125	10	3.25	1500	210	70	—	6.5	8.0	5.0	30	J.	3N	Class-C Amp. (Telegraphy)	1500	−105	200	40	8.5	—	215
														Class-C Amp. (Telephony)	1250	−160	160	60	16	—	140
														Class-B Amp. (Audio)[7]	1500	−16	84/400	280[9]	7.0[8]	8200	370
T125	125	10	4.5	2500	250	60	25	6.3	6.0	1.3	60	J.	2N	Class-C Amp. (Telegraphy)	2500	−200	240	31	11	—	475
														Class-C Amp. (Telephony)	2000	−215	200	28	10	—	320
HF130	125	10	3.25	1250	210	—	12.5	5.5	9.0	3.5	20	J.	—	Class-C Amp.-Oscillator	1250	−250	200	10	3.5	—	170
HF150	125	10	3.25	1500	210	—	12.5	5.5	7.2	1.9	30	J.	—	Class-C Amp.-Oscillator	1500	−300	200	10	4	—	220
HF175	125	10	4.0	2000	250	—	18	4.8	6.3	2.7	25	J.	T-3AC	Class-C Amp.-Oscillator	2000	−250	200	23	9	—	320

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Connec-tions	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx Output Power Watts
GL146	125	10	3.25	1500	200	60	75	7.2	9.2	3.9	15	J.	T-4BG	Class-C Amp.-Oscillator	1250	−150	180	30	—	—	150
														Class-C Amp. (Telephony)	1000	−200	160	40	—	—	100
														Class-B Amp. (Audio) [7]	1250	0	34/320	—	—	8400	250
GL152	125	10	3.25	1500	200	60	25	7.0	8.8	4.0	15	J.	T-4BG	Class-C Amp.-Oscillator	1250	−150	180	30	—	—	150
														Class-C Amp. (Telephony)	1000	−200	160	30	—	—	100
														Class-B Amp. (Audio) [7]	1250	− 40	16/320	—	—	8400	250
805	125	10	3.25	1500	210	70	40/60	8.5	6.5	10.5	30	J.	3N	Class-C Amp. (Telegraphy)	1500	−105	200	40	8.5	—	215
														Class-C Amp. (Telephony)	1250	−160	160	60	16	—	140
														Class-B Amp. (Audio) [7]	1500	− 16	84/400	280 [9]	7.0 [8]	8200	370
3X150A3 3C37	150	6.3	2.5	1000	—	—	23	4.2	3.5	0.6	500	N.	—	—	—	—	—	—	—	—	—
150T [1]	150	5.0	10	3000	200	50	13	3.0	3.5	0.5	—	J.	2N	Class-C Amp. (Telegraphy)	3000	−600	200	35	—	—	450
3-150A3 152TH	150	5/10	12.5/6.25	3000	450	85	20	5.7	4.5	0.8	40	J.	4BC	Class-C Amp. (Telegraphy)	3000	−300	250	70	27	—	600
														Class-B Amp. (Audio) [7]	3000	−150	67/335	430 [9]	3.0 [8]	20300	700
3-150A2 152TL						75	12	4.5	4.4	0.7			4BC	Class-C Amp. (Telegraphy)	3000	−400	250	40	20	—	600
														Class-B Amp. (Audio) [7]	3000	−260	65/335	675 [9]	3.0 [8]	20400	700
TW150	150	10	4.1	3000	200	60	35	3.9	2.0	0.8	—	J.	2N	Class-C Amp.-Oscillator	3000	−170	200	45	17	—	470
														Class-C Amp. (Telephony)	3000	−260	165	40	17	—	400
HK252-L	150	5/10	13/6.5	3000	500	75	10	7.0	5.0	0.4	125	N.	4BC	Class-C Amp.-Oscillator	3000	−400	250	30	15	—	610
														Class-C Amp. (Telephony)	2500	−350	250	35	16	—	500
DR200 HF200 HV18	150	10–11	3.4	2500	200	50	18	5.2	5.8	1.2	20	J.	2N	Class-C Amp. (Telegraphy)	2500	−300	200	18	8.0	—	380
														Class-C Amp. (Telephony)	2000	−350	160	20	9.0	—	250
														Class-B Amp. (Audio) [7]	2500	−130	60/360	460 [9]	8.0 [8]	16000	600
HD203A	150	10	4.0	2000	250	60	25	—	12	—	15	J.	3N	Class-C Amplifier	—	—	—	—	—	—	375
HF250	150	10.5	4.0	2500	200	—	18	—	5.8	—	20	J.	2N	Class-C Amp.-Oscillator	2500	—	200	—	—	—	375
HK354 HK354C	150	5.0	10	4000	300	50	14	4.5	3.8	1.1	30	J.	2N	Class-C Amp. (Telegraphy)	4000	−690	245	50	48	—	830
														Class-C Amp. (Telephony)	3000	−550	210	50	35	—	525
														Grid-Modulated Amp.	3000	−400	78	3.0	12	—	85
														Class-B Amp. (Audio) [7]	3000	−205	65/313	630 [9]	20 [8]	22000	665
HK354D	150	5.0	10	4000	300	55	22	4.5	3.8	1.1	30	J.	2N	Class-C Amp. (Telegraphy)	3500	−490	240	50	38	—	690
														Class-C Amp. (Telephony)	3500	−425	210	55	36	—	525
HK354E	150	5.0	10	4000	300	60	35	4.5	3.8	1.1	30	J.	2N	Class C Amp. (Telegraphy)	3500	−448	240	60	45	—	690
														Class-C Amp. (Telephony)	3000	−437	210	60	45	—	525
HK354F	150	5.0	10	4000	300	75	50	4.5	3.0	1.1	30	J.	2N	Class-C Amp. (Telegraphy)	3500	−368	250	75	50	—	720
														Class-C Amp. (Telephony)	3000	−312	210	75	45	—	525
UE-468	150	10	4.05	2500	200	60	18	8.8	7.0	1.25	30	J.	Fig. 57	Class-C Amp. (Telegraphy)	2500	−300	200	18	8.0	—	380
														Class-C Amp. (Telephony)	2000	−350	160	20	9.0	—	250
														Class-B (Audio) [7]	2500	−130	320 [8]	410 [9]	2.5	16000	500
810 1627 [1]	175	10 5.0	4.5 9.0	2500	300	75	36	8.7	4.8	12	30	J.	2N	Class-C Amp. (Telegraphy)	2500	−180	300	60	19	—	575
														Class-C Amp. (Telephony)	2000	−350	250	70	35	—	380
														Grid-Modulated Amp.	2250	−140	100	2.0	4.0	—	75
														Class-B Amp. (Audio) [7]	2250	− 60	70/450	380 [9]	13 [8]	11600	725

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amp.	Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Connections	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
8000	175	10	4.5	2500	300	45	16.5	5.0	6.4	3.3	30	J.	2N	Class-C Amp.-Oscillator	2500	−240	300	40	18	—	575
														Class-C Amp. (Telephony)	2000	−370	250	37	20	—	380
														Grid-Modulated Amp.	2250	−265	100	0	2.5	—	75
														Class-B Amp. (Audio) [7]	2250	−130	65/450	560 [9]	7.9 [8]	12000	725
GL-5C24	160	10	5.2	1750	107	—	8	5.6	8.8	3.3	—	N.	Fig. 26	Class-A Amp. (Audio)	1500	−155	107	—	—	8200 [5]	55
														Class-AB1 Amp. (Audio) [7]	1750	−200	320 [8]	390 [9]	—	8000	240
RK63 RK63A	200	5.0 6.3	10 14	3000	250	60	37	2.7	3.3	1.1	—	J.	2N	Class-C Amp. (Telegraphy)	3000	−200	233	45	17	—	525
														Class-C Amp. (Telephony)	2500	−200	205	50	19	—	405
														Grid-Modulated Amp.	3000	−250	100	7.0	12.5	—	100
T200	200	10	5.75	2500	350	80	16	9.5	7.9	1.6	30	J.	2N	Class-C Amp. (Telegraphy)	2500	−280	350	54	25	—	685
														Class-C Amp. (Telephony)	2000	−260	300	54	23	—	460
F-127-A	200	10	4.0	3000	325	70	38	13	4	13	—	J.	Fig. 26	Class-C Amp. (Telegraphy)	3000	−250	250	47	18	—	600
														Class-C Amp. (Telephony)	2500	−300	200	58	25.2	—	420
														Class-B Amp. (Audio) [7]	2800	− 75	20/400	175 [9]	6.65 [8]	16600	820
822 822S	200	10	4.0	2500	300	60	30	8.5	13.5	2.1	20 30	J.	3N 2N	Class-C Amp. (Telegraphy)	2500	−190	300	51	17	—	600
														Class-C Amp. (Telephony)	2000	− 75	250	43	13.7	—	405
														Class-B Amp. (Audio) [7]	3000	− 80	450 [8]	362 [9]	8.0 [8]	16000	1000
4C32	200	10	4.5	3000	300	60	30	5.5	5.8	1.1	60	J.	2N	Class-C Amp.-Oscillator	2000	−165	275	20	10	—	400
														Class-C Amp. (Telephony)	2000	−200	250	20	15	—	375
GL-592	200	10	5.0	3500	250	50	24	3.6	3.3	0.41	110	N.	Fig. 52	Class-C Amp.-Oscillator	2600	−240	250	45	18	—	425
														Class-C Amp. (Telephony)	2000	−500	250	50	—	—	—
4C34 HF300	200	11–12	4.0	3000	275	60	23	6.0	6.5	1.4	60 20	J.	2N	Class-C Amp. (Telegraphy)	3000	−400	250	28	16	—	600
														Class-C Amp. (Telephony)	2000	−300	250	36	17	—	385
														Class-B Amp. (Audio) [7]	3000	−115	60/360	450 [9]	13 [8]	20000	780
T814 HV12	200	10	4.0	2500	200	60	12	8.5	12.8	1.7	30	J.	3N	Class-C Amp. (Telegraphy)	2500	−240	300	30	10	—	575
														Class-C Amp. (Telephony)	2000	−370	300	40	20	—	485
														Class-B Amp. (Audio) [7]	2000	−160	50/275	350 [9]	7.0 [8]	14400	400
T822 HV27	200	10	4.0	2500	300	60	27	8.5	13.5	2.1	30	J.	3N	Class-C Amp. (Telegraphy)	2500	−175	300	50	15	—	585
														Class-C Amp. (Telephony)	2000	−195	250	45	15	—	400
T-300	200	11	6.0	3000	300	—	23	6.0	7.0	1.4	—	—	—	Class-C Amp. (Telegraphy)	3000	−400	250	28	20	—	600
														Class-C Amp. (Telephony)	2000	−300	250	36	17	—	385
														Class-B (Audio) [7]	2500	−100	60/450	—	7.5 [8]	—	750
806	225	5.0	10	3300	300	50	12.6	6.1	4.2	1.1	30	J.	2N	Class-C Amp. (Telegraphy)	3300	−600	300	40	34	—	780
														Class-C Amp. (Telephony)	3000	−670	195	27	24	—	460
														Class-B Amp. (Audio) [7]	3300	−240	80/475	930 [9]	35 [8]	16000	1120
3-250A4 250TH	250	5.0	10.5	4000	350	100	37	5.0	2.9	0.7	40	J.	2N	Class-C Amp. (Telegraphy)	2000	−120	350	100	34	—	500
														Class-C Amp. (Telephony)	3000	−210	330	75	42	—	750
														Grid-Modulated Amp.	3000	−160	125	4.5	20	—	125
														Class-B Amp. (Audio) [7]	3000	− 65	100/560	460 [9]	24 [8]	12250	1150
3-250A2 250TL	250	5.0	10.5	4000	350	50	14	3.7	3.1	0.7	40	J.	2N	Class-C Amp. (Telegraphy)	3000	−350	335	45	29	—	750
														Class-C Amp. (Telephony)	3000	−350	335	45	29	—	750
														Grid-Modulated Amp.	3000	−450	125	2.0	15	—	125
														Class-B Amp. (Audio) [7]	3000	−175	100/500	840 [9]	17 [8]	13000	1000

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Connec-tions	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
GL159	250	10	9.6	2000	400	100	20	11	17.6	5.0	15	J.	T-4BG	Class-C Amp.-Oscillator	2000	−200	400	17	6.0	—	620
														Class-C Amp. (Telephony)	1500	−240	400	23	9.0	—	450
														Class-B Amp. (Audio)[7]	2000	−100	30/660	400[9]	4.0[8]	6880	900
GL169	250	10	9.6	2000	400	100	85	11.5	19	4.7	15	J.	T-4BG	Class-C Amp.-Oscillator	2000	−100	400	42	10	—	620
														Class-C Amp. (Telephony)	1500	−100	400	45	10	—	450
														Class-B Amp. (Audio)[7]	2000	− 18	30/660	220[9]	6.0[8]	7000	900
204A 304A	250	11	3.85	2500	275	80	23	12.5	15	2.3	3	N.	T-1A	Class-C Amp. (Telegraphy)	2500	−200	250	30	15	—	450
														Class-C Amp. (Telephony)	2000	−250	250	35	20	—	350
														Class-B Amp. (Audio)[7]	3000	−100	80/372	500[9]	18[8]	20000	700
308B	250	14	4.0	2250	325	75	8.0	13.6	17.4	9.3	1.5	N.	T-2A	Class-C Amp. (Telegraphy)	1750	−345	300	—	—	—	350
															1500	−300	300	—	—	—	300
														Class-B Amp. (Audio)[7]	1750	−215	30/300	—	35[8]	5200	575
HK454H	250	5.0	11	5000	375	85	30	4.6	3.4	1.4	100	J.	2N	Class-C Amp. (Telegraphy)	3500	−275	270	60	28	—	760
HK454-L	250	5.0	11	5000	375	60	12	4.6	3.4	1.4	100	J.	2N	Class-C Amp. (Telephony)	3500	−450	270	45	30	—	760
212E 241B 312E	275	14	4.0	3000	350	75	16	14.9	18.8	8.6	1.5	N.	T-2A T-2AA	Class-C Amp. (Telegraphy)	3500	−275	270	60	28	—	760
														Class-C Amp. (Telephony)	3500	−450	270	45	30	—	760
														Class-B Amp. (Audio)[7]	2000	−105	40/300	—	50[8]	8000	650
300T[1]	300	8.0	11.5	3500	350	75	16	4.0	4.0	0.6	—	J.	2N	Class-C Amp. (Telegraphy)	2000	−225	300	—	—	—	400
HK304-L	300	5/10	26/13	3000	1000	150	10	12	9.0	0.8	—	N.	4BC	Class-C Amp. (Telephony)	1500	−200	300	75	—	—	300
527	300	5.5	135.0	—	—	—	38	19.0	12.0	1.4	200	N.	T-4B	Oscillator at 200 Mc.		Approximately 250 watts output					
HK654	300	7.5	15	4000	600	100	22	6.2	5.5	1.5	20	J.	2N	Class-C Amp. (Telegraphy)	2000	−380	500	75	57	—	720
														Class-C Amp. (Telephony)	2000	−365	450	110	70	—	655
														Grid-Modulated Amp.	3500	−210	150	15	15	—	210
3-300A3 304TH	300	5/10	25/12.5	3000	900	170	20	13.5	10.2	0.7	40	N.	4BC	Class-C Amplifier	1500	−125	667	115	25	—	700
														Class-B Amp. (Audio)[7]	3000	−150	134/667	420[9]	6.0[8]	10200	1400
3-300A2 304TL						150	12	8.5	9.1	0.6	40	N.	4BC	Class-C Amplifier	1500	−250	665	90	33	—	700
														Class-B Amp. (Audio)[7]	3000	−260	130/667	650[9]	6.0[8]	10200	1400
833A	350	10	10	3300	500	100	35	12.3	6.3	8.5	30	N.	T-1AB	Class-C Amp. (Telegraphy)	2000	−200	475	65	25	—	740
														Class-C Amp. (Telephony)	2500	−300	335	75	30	—	635
270A	350	10	4.0	3000	375	75	16	18	21	2.0	7.5	N.	T-1A	Class-C Amp. (Telegraphy)	3000	−375	350	—	—	—	700
														Class-C Amp. (Telephony)	2250	−300	300	80	—	—	450
849[1]	400	11	5.0	2500	350	125	19	17	33.5	3.0	3	N.	T-1A	Class-C Amp. (Telegraphy)	2500	−250	300	20	8.0	—	560
														Class-C Amp. (Telephony)	2000	−300	300	30	14	—	425
831[1]	400	11	10	3500	350	75	14.5	3.8	4.0	1.4	—	N.	T-1AA	Class-C Amp. (Telegraphy)	3500	−400	275	40	30	—	590
														Class-C Amp. (Telephony)	3000	−500	200	60	50	—	360

* Cathode resistor in ohms.

[1] Discontinued.
[2] Twin triode. Values, except interelement capacities, are for both sections in push-pull.
[3] Output at 112 Mc.
[4] Grid-leak resistor in ohms.
[5] Max. peak volts, plate pulsed.
[6] Per section.
[7] Values are for two tubes in push-pull.
[8] Max. signal value.
[9] Peak a.f. grid-to-grid volts.
[10] For single tube.
[11] Class-B data in Table I.
[12] Forced-air cooling.

TABLE XVII—TETRODE AND PENTODE TRANSMITTING TUBES

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	Max. Plate Volt-age	Max. Screen Volt-age	Max. Screen Dissi-pation Watts	Interelectrode Capacitances ($\mu\mu$fd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Con-nec-tions	Typical Operation	Plate Volt-age	Screen Volt-age	Sup-pressor Volt-age	Grid Volt-age	Plate Current Ma.	Screen Current Ma.	Grid Current Ma.	Screen Resistor Ohms	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
3A4	2.0	1.4 2.0	0.2 0.1	150	135	0.9	4.8	0.2	4.2	10	B.	7BB	Class-C Amp. (Telegraphy)	150	135	0	−26	18.3	6.5	0.13	2300	—	—	1.2
3D6	—	2.8 1.4	0.11 0.22	150	135	—	7.5	0.3	6.5	50	L.	6BB	Class-C Amp. (Telegraphy)	150	135	—	−20	23	6.0	—	—	—	—	1.4
3B4	3.0	2.5 1.25	0.165 0.33	150	135	—	4.6	0.16	7.6	100	B.	7CY	Class-C Amp.	150	135	—	−75	25	—	—	—	—	—	1.25
HY63[1]	3.0	2.5 1.25	0.1125 0.225	200	100	0.6	8.0	0.1	8.0	60	O.	T-8DB	Class-C Amp. (Telegraphy)	200	100	—	−22.5	20	4.0	2.0	—	0.1	—	3.0
													Class-C Amp. (Telephony)	180	100	—	−35	15	3.0	2.0	—	0.2	—	2.0
6AK6	3.5	6.3	0.15	375	250	1.0	3.6	0.12	4.2	54	B.	7BK	Class-C Amp. (Telegraphy)	375	250	—	−100	13	4.0	3.0	—	—	—	4.0
5A6	5	2.5 5.0	0.46 0.23	150	150	2	8.5	0.15	9.5	100	B.	9L	Class-C Amp.	150	150	0	−24	40	11	1.2	—	—	—	3.1
5618	5.0	6.0 3.0	0.23 0.46	300	125	2.0	7.0	0.24	5.0	80	B.	7CU	Class-C Amp. (Telegraphy)	300	75	0	−45	25	7.0	1.5	32000	0.3	—	5.4
6AQ5	8.0	6.3	0.45	350	250	2.0	7.6	0.35	6.0	54	B.	7BZ	Class-C Amp. (Telegraphy)	350	250	—	−100	47	7.0	5.0	—	—	—	11
6V6GT	8.0	6.3	0.45	350	250	2.0	9.5	0.7	7.5	10	O.	7AC	Class-C Amp. (Telegraphy)	350	250	—	−100	47	7.0	5.0	—	—	—	11
6AG7	9.0	6.3	0.65	375	250	1.5	13	0.06	7.5	10	O.	8Y	Class-C Amp. (Telegraphy)	375	250	—	−75	30	9.0	5.0	—	—	—	7.5
RK64[1]	6.0	6.3	0.5	400	100	3.0	10	0.4	9.0	60	M.	5AW	Class-C Amp. (Telegraphy)	400	100	30	−30	35	10	3.0	—	0.18	—	10
													Class-C Amp. (Telephony)	300	—	30	−30	26	8.0	1.0	30000	0.2	—	6.0
1610	6.0	2.5	1.75	400	200	2.0	8.6	1.2	13	20	M.	T-5CA	Class-C Amp. (Telegraphy)	400	150	—	−50	22.5	7.0	1.5	—	0.1	—	5.0
RK56	8.0	6.3	0.55	300	300	4.5	10	0.2	9.0	60	M.	5AW	Class-C Amp. (Telegraphy)	400	300	—	−40	62	12	1.6	—	0.1	—	12.5
													Class-C Amp. (Telephony)	250	200	—	−40	50	10	1.6	2800	0.28	—	8.5
RK23[1] RK25 RK25B[1]	10	2.5 6.3	2.0 0.9	500	250	8	10	0.2	10	—	M.	6BM	Class-C Amp. (Telegraphy)	500	200	45	−90	55	38	4.0	—	0.5	—	22
													Class-C Amp. (Telephony)	400	150	0	−90	43	30	6.0	8300	0.8	—	13.5
													Suppressor-Modulated Amp.	500	200	−45	−90	31	39	4.0	—	0.5	—	6.0
1613	10	6.3	0.7	350	275	2.5	8.5	0.5	11.5	45	O	7S	Class-C Amp. (Telegraphy)	350	200	—	−35	50	10	3.5	20000	0.22	—	9
													Class-C Amp. (Telephony)	275	200	—	−35	42	10	2.8	10000	0.16	—	6.0
2E30	10	6.0	0.7	250	250	2.5	10	0.5	4.5	160	B.	7CQ	Class-C Amp. (Telegraphy)	250	200	—	−50	50	10	2.5	—	0.2	—	7.5
													Class-AB$_2$ Amp. (Audio)[6]	250	250	—	−30	40/120	4/20	2.3[7]	87[8]	0.2	3800	17
837 RK44[1]	12	12.6	0.7	500	300	8	16	0.2	10	20	M.	6BM	Class-C Amp. (Telegraphy)	500	200	40	−70	80	15	4.0	20000	0.4	—	28
													Class-C Amp. (Telephony)	400	140	40	−40	45	20	5.0	13000	0.3	—	11
													Suppressor-Modulated Amp.	500	—	−65	−20	30	23	3.5	14000	0.1	—	5.0
5763	12	6.0	0.75	300	250	2	9.5	0.3	4.5	175	B.	9K	Class-C Amp. (Telegraphy)	300	250	0	−60	50	5.0	3.0	—	0.35	—	8.0
													Doubler to 175 Mc.	300	250	0	−75	40	4.0	1.0	12500	0.6	—	3.6
6F6 6F6G	12.5	6.3	0.7	400	275	3.0	6.5 8.0	0.2 0.5	13 6.5	10	O.	7AC	Class-C Amp. (Telegraphy)	400	275	—	−100	50	11	5.0	—	—	—	14
													Class-C Amp. (Telephony)	275	200	—	−35	42	10	2.8	—	0.16	—	6.0
2E24	9.0	6.3[5]	0.65	500	200	2.3	8.5	0.11	6.5	125	O.	7CL	Class-C Amp. (Telephony)	400	180	—	−45	50	8.0	2.5	27500	0.15	—	13.5
														500	180	—	−45	54	8.0	2.5	40000	0.16	—	18.0
	13.5			600	200	2.5							Class-C Amp. (Telegraphy)	400	200	—	−45	75	10.0	3.0	20000	0.19	—	20
														600	195	—	−50	66	10	3.0	40500	0.21	—	27
2E26	13.5	6.3	0.8	600	200	2.5	13	0.2	7.0	125	O.	7CK	Class-C Amp. (Telegraphy)	600	185	—	−45	66	10	3.0	41500	0.17	—	27
	9.0			500	200	2.3							Class-C Amp. (Telephony)	500	180	—	−50	54	9.0	2.5	35500	0.15	—	18
													Class-AB$_2$ Amp. (Audio)[6]	500	125	—	−15	22/150	32[7]	—	60[8]	0.36[7]	8000	54
802	13	6.3	0.9	600	250	6.0	12	0.15	8.5	30	M.	6BM	Class-C Amp. (Telegraphy)	600	250	40	−120	55	16	2.4	22000	0.30	—	23
													Class-C Amp. (Telephony)	500	245	40	−40	40	15	1.5	16300	0.10	—	12
													Suppressor-Modulated Amp.	600	250	−45	−100	30	24	5.0	14500	0.6	—	6.3
HY6V6-GTX	13	6.3	0.5	350	225	2.5	9.5	0.7	9.5	60	O.	7AC	Class-C Amp. (Telegraphy)	300	200	—	−45	60	7.5	2.5	—	0.3	—	12
													Class-C Amp. (Telephony)	250	200	—	−45	60	6.0	2.0	15000	0.4	—	10

TABLE XVII—TETRODE AND PENTODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	Max. Plate Volt-age	Max. Screen Volt-age	Max. Screen Dissi-pation Watts	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Con-nec-tions	Typical Operation	Plate Volt-age	Screen Volt-age	Sup-pressor Volt-age	Grid Volt-age	Plate Current Ma.	Screen Current Ma.	Grid Current Ma.	Screen Resistor Ohms	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
HY60	15	6.3	0.5	425	225	2.5	10	0.2	8.5	60	M.	5AW	Class-C Amp. (Telegraphy)	425	200	—	−62.5	60	8.5	3.0	—	0.3	—	18
													Class-C Amp. (Telephony)	325	200	—	−45	60	7.0	2.5	—	0.2	—	14
HY65[1]	15	6.3	0.85	450	250	4.0	9.1	0.18	7.2	60	O.	T-8DB	Class-C Amp.-Oscillator	450	250	—	−45	75	15	3.0	—	0.5	—	24
													Class-C Amp. (Telephony)	350	200	—	−45	63	12	3.0	—	0.5	—	16
2E25	15	6.0	0.8	450	250	4.0	8.5	0.15	6.7	125	O.	5BJ	Class-C Amp.-Oscillator	450	250	—	−45	75	15	3.0	—	0.4	—	24
													Class-C Amp. (Telephony)	400	200	—	−45	60	12	3.0	—	0.4	—	16
													Class-AB$_2$ Amp. (Audio)[6]	450	250	—	−30	44/150	10/40	3.0	142[8]	0.9[7]	6000	40
306A	15	2.75	2.0	300	300	6.0	13	0.35	13	—	M.	T-5CB	Class-C Amp. (Telephony)	300	180	—	−50	36	15	3.0	8000	—	—	7.0
307A RK-75	15	5.5	1.0	500	250	6.0	15	0.55	12	—	M.	T-5C	Class-C Amp. (Telegraphy)	500	250	0	−35	60	13	1.4	20000	—	—	20
													Suppressor-Modulated Amp.	500	200	−50	−35	40	20	1.5	14000	—	—	6.0
832[3]	15	6.3 12.6	1.6 0.8	500	250	5.0	7.5	0.05	3.8	200	N.	7BP	Class-C Amp. (Telegraphy)	500	200	—	−65	72	14	2.6	21000	0.18	—	26
													Class-C Amp. (Telephony)	425	200	—	−60	52	16	2.4	14000	0.15	—	16
832A[3]	15	6.3 12.6	1.6 0.8	750	250	5.0	7.5	0.05	3.8	200	N.	7BP	Class-C Amp. (Telegraphy)	750	200	—	−65	48	15	2.8	36500	0.19	—	26
													Class-C Amp. (Telephony)	600	200	—	−65	36	16	2.6	25000	0.16	—	17
844[1]	15	2.5	2.5	500	180	3.0	9.5	0.15	7.5	—	M.	5AW	Class-C Amp. (Telegraphy)	500	175	—	−125	25	—	5.0	—	—	—	9.0
													Class-C Amp. (Telephony)	500	150	—	−100	20	—	—	—	—	—	4.0
865	15	7.5	2.0	750	175	3.0	8.5	0.1	8.0	15	M.	T-4C	Class-C Amp. (Telegraphy)	750	125	—	−80	40	—	5.5	—	1.0	—	16
													Class-C Amp. (Telephony)	500	125	—	−120	40	—	9.0	—	2.5	—	10
1619	15	2.5	2.0	400	300	3.5	10.5	0.35	12.5	45	O.	T9H	Class-C Amp. (Telegraphy)	400	300	—	−55	75	10.5	5.0	9500	0.36	—	19.5
													Class-C Amp. (Telephony)	325	285	—	−50	62	7.5	2.8	5000	0.18	—	13
													Class-AB$_2$ Amp. (Audio)[6]	400	300	0	−16.5	75/150	6.5/11.5	—	77[8]	0.4[7]	6000	36
5516	15	6.0	0.7	600	250	5.0	8.5	0.12	6.5	80	O.	7CL	Class-C Amp. (Telegraphy)	600	250	—	−60	75	15	5.0	—	0.5	—	32
													Class-C Amp. (Telephony)	475	250	—	−90	63	10	4.0	22500	0.5	—	22
													Class-AB$_2$ (Audio)[6]	600	250	—	−25	36/140	1/24	4[7]	80[8]	0.16	10500	67
254A	20	5.0	3.25	750	175	5.0	4.6	0.1	9.4	—	M.	T-4C	Class-C Amplifier	750	175	—	−90	60	—	—	—	—	—	25
6L6 6L6G	21	6.3	0.9	400	300	3.5	10 11.5	0.4 0.9	12 9.5	10	O.	7AC	Class-C Amp.-Oscillator	400	300	—	−125	100	12	5.0	—	—	—	28
													Class-C Amp. (Telephony)	325	250	—	−70	65	—	9.0	—	0.8	—	11
6L6GX	21	6.3	0.9	500	300	3.5	11	1.5	7.0	—	O.	7AC	Class-C Amp. (Telegraphy)	500	250	—	−50	90	9.0	2.0	—	0.25	—	30
													Class-C Amp. (Telephony)	325	225	—	−45	90	9.0	3.0	—	0.25	—	20
HY6L6-GTX	21	6.3	0.9	500	300	3.5	11	0.5	7.0	60	O.	7AC	Class-C Amp.-Oscillator	500	250	—	−50	90	9.0	2.0	—	0.5	—	30
													Class-C Amp. (Telephony)	400	225	—	−45	90	9.0	3.0	16000	0.8	—	20
T21	21	6.3	0.9	400	300	3.5	13	0.7	12	30	M.	6A	Class-C Amp. (Telegraphy)	400	250	—	−50	95	8.0	3.0	—	0.2	—	25
													Class-C Amp. (Telephony)	350	200	—	−45	65	17	5.0	—	0.35	—	14
RK49	21	6.3	0.9	400	300	3.5	11.5	1.4	10.6	—	M.	6A	Class-C Amp. (Telegraphy)	400	250	—	−50	95	8.0	3.0	—	0.2	—	25
													Class-C Amp. (Telephony)	300	200	—	−45	60	15	5.0	6700	0.34	—	12
1614	25	6.3	0.9	450	300	3.5	10	0.4	12.5	80	O.	7AC	Class-C Amp. (Telegraphy)	450	250	—	−45	100	8	2.0	12500	0.15	—	31
													Class-C Amp. (Telephony)	375	250	—	−50	93	7.0	2.0	10000	0.15	—	24.5
													Class-AB$_1$ Amp. (Audio)[6]	530	340	—	−36	60/160	20[7]	—	72[8]	—	7200	50
RK41[1] RK39	25	2.5 6.3	2.4 0.9	600	300	3.5	13	0.2	10	30	M.	5AW	Class-C Amp. (Telegraphy)	600	300	—	−90	93	10	3.0	—	0.38	—	36
													Class-C Amp. (Telephony)	475	250	—	−50	85	9.0	2.5	25000	0.2	—	26
HY61/807	25	6.3	0.9	600	300	3.5	11	0.2	7.0	60	M.	5AW	Class-C Amp. (Telegraphy)	600	250	—	−50	85	9.0	4.0	39000	0.4	—	40
													Class-C Amp. (Telephony)	475	250	—	−50	100	9.0	3.5	25000	0.2	—	27
													Class-AB$_2$ Amp. (Audio)[6]	600	300	—	−30	200[7]	10[7]	—	—	0.1[7]	—	80
815[3]	25	12.6 6.3	0.8 1.6	500	200	4.0	13.3	0.2	8.5	125	O.	8BY	Class-C Amp.-Oscillator	500	200	—	−45	150	17	2.5	—	0.13	—	56
													Class-C Amp. (Telephony)	400	175	—	−45	150	15	3.0	—	0.16	—	45
													Class-AB$_2$ Amp. (Audio)[3]	500	125	—	−15	22/150	32[7]	—	60[8]	0.36[7]	8000	54

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	Max. Plate Volt-age	Max. Screen Volt-age	Max. Screen Dissi-pation Watts	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Con-nec-tions	Typical Operation	Plate Volt-age	Screen Volt-age	Sup-pressor Volt-age	Grid Volt-age	Plate Current Ma.	Screen Current Ma.	Grid Current Ma.	Screen Resistor Ohms	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
254B	25	7.5	3.25	750	150	5.0	11.2	0.085	5.4	—	M.	T-4C	Class-C Amplifier	750	150	—	−135	75	—	—	—	—	—	30
1624	25	2.5	2.0	600	300	3.5	11	0.25	7.5	60	M.	T-5DC	Class-C Amp. (Telegraphy)	600	300	—	−60	90	10	5.0	30000	0.43	—	35
													Class-C Amp. (Telephony)	500	275	—	−50	75	9.0	3.3	25000	0.25	—	24
													Class-AB$_2$ Amp. (Audio)[6]	600	300	—	−25	42/180	5/15	106[8]	—	1.2[7]	7500	72
3DX3	25	6.3	3.0	1500	200	—	—	—	—	250	S.	Fig. 40	Class-C Amp. (Telegraphy)	1000	200	—	−155	75	—	2.8	—	0.57	—	50
3E22[3]	30	12.6 6.3	0.8 1.6	560	225	6.0	14	0.22	8.5	200	O.	8BY	Class-C Amp. (Telegraphy)[3]	600	200	—	−55	160	20	7.0	20000	0.45	—	72
													Class-C Amp. (Telephony)[3]	560	200	—	−50	160	20	6.5	18000	0.4	—	67
RK66	30	6.3	1.5	600	300	3.5	12	0.25	10.5	60	M.	T-5C	Class-C Amp.-Oscillator	600	300	—	−60	90	11	5.0	—	0.5	—	40
													Class-C Amp. (Telephony)	500	—	—	−50	75	8.0	3.2	25000	0.23	—	25
807 1625	30	6.3 12.6	0.9 0.45	750	300	3.5	11	0.2	7.0	60	M.	5AW 5AZ	Class-C Amp. (Telegraphy)	750	250	—	−45	100	6	3.5	85000	0.22	—	50
													Class-C Amp. (Telephony)	600	275	—	−90	100	6.5	4.0	50000	0.4	—	42.5
													Class-AB$_2$ Amp. (Audio)[6]	750	300	—	−32	60/240	5/10	92[8]	—	0.2[7]	6950	120
													Class-B Amp. (Audio)[11]	750	—	—	0	15/240	—	555[b]	—	5.3[7]	6650	120
2E22	30	6.3	1.5	750	250	10	13	0.2	8.0	—	M.	5J	Class-C Amp.-Oscillator	500	250	22.5	−60	100	16	6.0	15000	0.55	—	34
													Class-C Amp.-Oscillator	750	250	22.5	−60	100	16	6.0	30000	0.55	—	53
													Suppressor-Modulated Amp.	750	250	−90	−65	55	29	6.5	17000	0.6	—	16.5
3D23 TB-35	35	6.3	3.0	—	—	—	6.5	0.2	1.8	250	M.	Fig. 54	Class-C Amp. (Telegraphy)	1500	375	—	−300	110	22	15	—	4.5	—	130
													Class-C Amp. (Telephony)	1000	300	—	−200	85	14	10	—	2.0	—	60
RK20[1] RK20A RK46[1]	40	7.5 7.5 12.6	3.0 3.25 2.5	1250	300	15	14	0.01	12	—	M.	T-5C	Class-C Amp. (Telegraphy)	1250	300	45	−100	92	36	11.5	—	1.6	—	84
													Class-C Amp. (Telephony)	1000	300	0	−100	75	30	10	23000	1.3	—	52
													Suppressor-Modulated Amp.	1250	300	−45	−100	48	44	11.5	—	1.5	—	21
													Grid-Modulated Amp.	1250	300	45	−142	40	7.0	1.8	—	1.5	—	20
HY69	40	6.3	1.5	600	300	5.0	15.4	0.23	6.5	60	M.	T-5D	Class-C Amp.-Oscillator	600	250	—	−60	100	12.5	4.0	30000	0.25	—	42
													Class-C Amp. (Telephony)	600	250	—	−60	100	12.5	5.0	30000	0.35	—	42
													Modulated Doubler	600	200	—	−300	90	11.5	6.0	35000	2.8	—	27
													Class-AB$_2$ Amp. (Audio)[6]	600	300	—	−35	200[7]	18[7]	5.0[7]	—	0.3[7]	—	60
829[1,3]	40	6.3 12.6	2.25 1.12	500	225	40	14.5	0.1	7.0	200	N.	7BP	Class-C Amp. (Telegraphy)	500	200	—	−45	240	32	12	9300	0.7	—	83
													Class-C Amp. (Telephony)	425	200	—	−60	212	35	11	6400	0.8	—	63
													Grid-Modulated Amp.	500	200	—	−38	120	10	2.0	—	0.5	—	23
829A[1,3]	40	6.3 12.6	2.25 1.12	750	240	7.0	14.4	0.1	7.0	200	N.	7BP	Class-C Amp.-Oscillator	750	200	—	−55	160	30	12	18300	0.8	—	87
													Class-C Amp. (Telephony)	600	200	—	−70	150	30	12	13300	0.9	—	70
													Grid-Modulated Amp.	750	200	—	−55	80	5.0	0	—	0.7	—	24
829B[3] 3E29[3]	30 28 40	12.6 6.3	1.125 2.25	750 600 750	225 225 240	6 7 7	14.5	0.12	7.0	200	N.	7BP	Class-C Amp. (Grid Mod.)	500	200	—	−38	120	10	2	—	0.5	—	23
													Class-C Amp. (Telephony)	425	200	—	−60	212	35	11.0	6400	0.8	—	63
													Class-C Amp. (Telegraphy)	500	200	—	−45	240	32	12.0	9300	0.7	—	83
HY1269	40	6.3 12.6	3.5 1.75	750	300	5.0	16.0	0.25	7.5	6	M.	T-5DB	Class-C Amp.-Oscillator	750	300	—	−70	120	15	4	—	0.25	—	63
													Class-C Amp. (Telephony)	600	250	—	−70	100	12.5	5	35000	0.5	—	42
													Grid-Modulated Amp.	750	300	—	—	80	—	—	—	—	—	20
													Class-AB$_2$ Amp. (Audio)[6]	600	300	—	−35	200[7]	—	—	—	0.3	—	80
3D24	45	6.3	3.0	2000	400	10	6.5	0.2	2.4	125	L.	T-9J	Class-C Amp.-Oscillator	2000	375	—	−300	90	20	10	—	4.0	—	140
														1500	375	—	−300	90	22	10	—	4.0	—	105
715-B	50	26/28	—	—	—	—	—	—	—	—	—	—	Class-C Amp. (Telegraphy)	1500	300	—	—	125	—	—	—	—	—	—
5562	45	6.3	3.0	2000	400	8	6.5	0.2	1.8	120	M.	Fig. 54	Class-C Amp. (Telegraphy)	1500	375	—	−300	116	21	12	—	3.6	—	135
													Class-C Amp. (Telephony)	1000	300	—	−200	85	14	10	—	2.0	—	60
HK-57	50	5	5	3000	500	25	7.29	0.05	3.13	200	N.	5BK	Class-C Amp. (Telegraphy)	2000	450	+30	−145	110	2	1	—	0.15	—	166
													Class-C Amp. (Telephony)	2000	450	+30	−145	88	2	1.5	—	0.2	—	135
													Suppressor-Modulated Amp.	2000	450	−190	−240	80	14	2.5	110000	0.6	—	90

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	Max. Plate Volt-age	Max. Screen Volt-age	Max. Screen Dissi-pation Watts	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Con-nec-tions	Typical Operation	Plate Volt-age	Screen Volt-age	Sup-pressor Volt-age	Grid Volt-age	Plate Current Ma.	Screen Current Ma.	Grid Current Ma.	Screen Resistor Ohms	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
RK47	50	10	3.25	1250	300	10	13	0.12	10	—	M.	T-5D	Class-C Amp. (Telegraphy)	1250	300	—	−70	138	14	7.0	—	1.0	—	120
													Class-C Amp. (Telephony)	900	300	—	−150	120	17.5	6.0	—	1.4	—	87
													Grid-Modulated Amp.	1250	300	—	−30	60	2.0	0.9	—	4.0	—	25
312A	50	10	2.0	1250	500	20	15.5	0.15	12.3	—	M.	T-6C	Class-C Amp. (Telegraphy)	1250	300	20	−55	100	36	5.5	—	0.7	—	90
													Class-C Amp. (Telephony)	1000	—	40	−40	95	35	7.0	22000	1.0	—	65
													Suppressor-Modulated Amp.	1250	—	−85	−50	50	42	5.0	22000	0.55	—	23
804	50	7.5	3.0	1500	300	15	16	0.01	14.5	15	M.	T-5C	Class-C Amp. (Telegraphy)	1500	300	45	−100	100	35	7.0	34000	1.95	—	110
													Class-C Amp. (Telephony)	1250	250	50	−90	75	20	6.0	50000	0.75	—	65
													Grid-Modulated Amp.	1500	300	45	−130	50	13.5	3.7	—	1.3	—	28
													Suppressor-Modulated Amp.	1500	300	−50	−115	50	32	7.0	—	0.95	—	28
4D22	50	25.2 12.6	0.8 1.6	750	350	14	28	0.27	13	60	N.	Fig. 50	Class-C Amp. (Telegraphy)	750	300	—	−100	240	26	12	—	1.5	—	135
														600	300	—	−100	215	30	10	—	1.25	—	100
4D32		6.3	3.75									Fig. 51	Class-C Amp. (Telephony)	600	—	—	−100	220	28	10	10000	1.25	—	100
														550	—	—	−100	175	17	6	15000	0.6	—	70
													Class-AB$_2$ Amp. (Audio) [6]	600	250	—	−25	100/365	26 [7]	70 [8]	—	0.45 [7]	3000	125
305A	60	10	3.1	1000	200	6	10.5	0.14	5.4	—	M.	T-4CE	Class-C Amp. (Telegraphy)	1000	200	—	−200	125	—	—	—	—	—	85
													Class-C Amp. (Telephony)	800	200	—	−270	125	—	—	—	—	—	70
HY67	65	6.3 12.6	4.5 2.25	1250	300	10	—	0.19	14.5	—	M.	T-5DB	Class-C Amp. (Telegraphy)	1250	300	—	−80	175	22.5	10	—	1.5	—	152
													Class-C Amp. (Telephony)	1000	300	—	−150	145	17.5	14	—	2.0	—	101
													Grid-Modulated Amp.	1250	300	—	—	78	—	—	—	—	—	32.5
814	65	10	3.25	1500	300	10	13.5	0.1	13.5	30	M.	T-5D	Class-C Amp. (Telegraphy)	1500	300	—	−90	150	24	10	50000	1.5	—	160
													Class-C Amp. (Telephony)	1250	300	—	−150	145	20	10	48000	3.2	—	130
													Grid-Modulated Amp.	1500	250	—	−120	60	3.0	2.5	—	4.2	—	35
4-65A	65	6.0	3.5	3000	400	10	8.0	0.08	2.1	160 [9]	N.	5BK	Class-C Amp. (Telegraphy)	3000	250	—	−90	115	20	10	—	1.7	—	280
				2500	400								Class-C Amp. (Telephony)	2500	250	—	−150	108	16	8	—	1.9	—	225
				3000	600								Class-B Linear Amp.	2500	500	—	−100	20/230	0/35	6 [10]	—	1.8 [10]	—	325 [7]
				3000	600								Class-AB$_2$ Amp. (Audio) [6]	1800	250	—	−35	50/220	0/25	180 [8]	—	2.2 [7]	20000	270
282A	70	10	3.0	1000	250	5	12.2	0.2	6.8	—	M.	T-4C	Class-C Amp. (Telegraphy)	1000	150	—	−160	100	—	—	—	—	—	33
													Class-C Amp. (Telephony)	750	150	—	−180	100	—	50	—	—	—	50
4E27/ 8001	75	5.0	7.5	4000	750	30	12	0.06	6.5	75	J.	T-7CB	Class-C Amp. (Telegraphy)	2000	500	60	−200	150	11	6	136000	1.4	—	230
													Class-C Amp. (Telephony)	1800	400	60	−130	135	11	8	125000	1.7	—	178
													Suppressor-Modulated Amp.	2000	500	−300	−130	55	27	3.0	—	0.4	—	35
HK257 HK257B	75	5.0	7.5	4000	750	25	13.8	0.04	6.7	75 120	J.	T-7CB	Class-C Amp. (Telegraphy)	2000	500	60	−200	150	11	6.0	—	1.4	—	230
													Class-C Amp. (Telephony)	1800	400	60	−130	135	11	8.0	—	1.7	—	178
													Suppressor-Modulated Amp.	2000	500	−300	−130	55	27	3.0	—	0.4	—	35
828	80	10	3.25	2000	750	23	13.5	0.05	14.5	30	M.	5J	Class-C Amp. (Telegraphy)	1500	400	75	−100	180	28	12	40000	2.2	—	200
													Class-C Amp. (Telephony)	1250	400	75	−140	160	28	12	30000	2.7	—	150
													Grid-Modulated Amp.	1500	400	75	−150	80	4.0	1.3	—	1.3	—	41
													Class-AB$_1$ Amp. (Audio) [6]	2000	750	60	−120	50/270	2/60	240 [3]	—	0	18500	385
RK28	100	10	5.0	2000	400	35	15	0.02	15	—	J.	5J	Class-C Amp. (Telegraphy)	2000	400	45	−100	150	55	13	21000	2.0	—	210
													Class-C Amp. (Telephony)	1500	400	45	−100	135	52	13	21000	2.0	—	155
													Suppressor-Modulated Amp.	2000	400	−45	−100	85	65	13	—	1.8	—	60
													Grid-Modulated Amplifier	2000	400	45	−140	80	20	4.0	—	0.9	—	75
RK48 RK48A	100	10	5.0	2000	400	22	17	0.13	13	—	J.	T-5D	Class-C Amp. (Telegraphy)	2000	400	—	−100	180	40	6.5	—	1.0	—	250
													Class-C Amp. (Telephony)	1500	400	—	−100	148	50	6.5	22000	1.0	—	165
													Grid-Modulated Amplifier	1500	400	—	−145	77	10	1.5	—	1.6	—	40

TABLE XVII—TETRODE AND PENTODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amp.	Max. Plate Voltage	Max. Screen Voltage	Max. Screen Dissipation Watts	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Connections	Typical Operation	Plate Voltage	Screen Voltage	Suppressor Voltage	Grid Voltage	Plate Current Ma.	Screen Current Ma.	Grid Current Ma.	Screen Resistor Ohms	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
813	100	10	5.0	2250	400	22	16.3	0.2	14	30	J.	5BA	Class-C Amp. (Telegraphy)	2250	400	0	−155	220	40	15	46000	4.0	—	375
													Class-C Amp. (Telephony)	2000	350	0	−175	200	40	16	41000	4.3	—	300
													Grid-Modulated Amplifier	2000	400	—	−120	75	3.0	—	—	—	—	50
													Class-B Amp. (Audio) [6]	2500	750	0	− 95	35/360	1.2/55	—	—	0.35	17000	650
850	100	10	3.25	1250	175	10	17	0.25	25	15	J.	T-38	Class-C Amp. (Telegraphy)	1250	175	—	−150	160	—	35	—	10	—	130
													Class-C Amp. (Telephony)	1000	140	—	−100	125	—	40	—	10	—	65
													Grid-Modulated Amplifier	1250	175	—	− 13	110	—	—	—	—	—	40
860	100	10	3.25	3000	500	10	7.75	0.08	7.5	30	M.	T-4CB	Class-C Amp.-Oscillator	3000	300	—	−150	85	25	15	—	7.0	—	165
													Class-C Amp. (Telephony)	2000	220	—	−200	85	25	38	100000	17	—	105
4-125A 4D21	125	5.0	6.2	3000	400	20	10.3	0.03	3.0	120	N.	5BK	Class-C Amp. (Telegraphy)	3000	350	—	−150	167	30	9	—	2.5	—	375
													Class-C Amp. (Telephony)	2500	350	—	−210	152	30	9	—	3.3	—	300
													Class-AB_2 Amp. (Audio) [6]	2500	350	—	− 43	93/260	0/6	178[8]	—	1.0	22200	400
RK28A	125	10	5.0	2000	400	35	15	0.02	15	—	J.	5J	Class-C Amp. (Telegraphy)	2000	400	45	−100	170	60	10	—	1.6	—	250
													Class-C Amp. (Telegraphy)	1500	400	45	−100	135	54	10	18500	1.6	—	150
													Grid-Modulated Amp.	2000	400	45	− 55	80	18	2.0	—	0.5	—	60
													Suppressor-Modulated Amp.	2000	—	−45	−115	90	52	11.5	30000	1.5	—	60
803	125	10	5.0	2000	600	30	17.5	0.15	29	20	J.	5J	Class-C Amp. (Telegraphy)	2000	500	40	− 90	160	45	12	—	2.0	—	210
													Class-C Amp. (Telephony)	1600	400	100	− 80	150	45	25	27000	5.0	—	155
													Suppressor-Modulated Amp.	2000	—	−110	−100	80	48	15	35000	2.5	—	53
													Grid-Modulated Amplifier	2000	600	40	− 80	80	20	4.0	—	2.0	—	53
4X-150A [9]	150	6.0	2.8	1000	300	15	16.1	0.02	4.7	500	N.	T-9J	Class-C Amp. (Telegraphy)	1000	250	—	− 80	200	39	7	—	0.69	—	148
														750	250	—	− 80	200	37	6.5	—	0.63	—	110
														600	250	—	− 75	200	35	6	—	0.52	—	85
PE340/ 4D23 [9]	150	5.0	7.5	4000	400	—	11.6	0.06	4.35	120	N.	5BK	Class-C Amp. (Telegraphy)	3000	400	—	−290	200	27	7	—	2.6	—	450
													Class-C Amp. (Telephony)	2500	400	—	−425	180	27	9	—	4	—	350
													Class AB_2 Audio [6]	2500	400	—	− 95	284[7]	7[7]	—	—	1.8[7]	19100	460
AT-340	150	5	7.0	4000	400	—	9.04	0.19	4.16	120	J.	5BK	Class-C Amp.-Oscillator	3000	400	—	−500	165	75	—	—	2.4	—	—
RK65	215	5.0	14	3000	500	35	10.5	0.24	4.75	60	J.	T-3BC	Class-C Amp. (Telegraphy)	3000	400	—	−100	240	70	24	—	6.0	—	510
													Class-C Amp. (Telephony)	2500	—	—	−150	200	70	22	30000	6.3	—	380
4-250A 5D22	250	5.0	14.5	4000	600	35	12.7	0.06	4.5	75	N.	5BK	Class-C Amp. (Telegraphy)	3000	500	—	−180	330	60	10	—	2.6	—	800
													Class-C Amp. (Telephony)	3000	400	—	−310	225	30	9	—	3.2	—	510
													Class-AB_2 (Audio) [6]	1500	300	—	− 48	100/485	0/34	192[8]	—	4.7[7]	5400	428
4-250A	250	5.0	14.5	4000	600	50	12.7	0.06	4.5	85	N.	5BK	Class-C Amp. (Telegraphy)	4000	500	—	−250	250	22	13	—	4.1	—	750
														2500	500	—	−100	325	70	22	—	3.7	—	562
GL-5D24	250	5.0	14.1	4000	350	50	12.7	0.06	4.5	85	N.	5BK	Class-C Amp. (Telegraphy)	Same as 4-250A										GL-5D24
4-400A [9]	400	5.0	14.5	4000	600	35	12.5	0.12	4.7	110	N.	5BK	Class-C Teleg. or Telephony	4000	300	—	−170	270	22.5	10	—	10	—	720
861	400	11	10	3500	750	35	14.5	0.1	10.5	20	N.	T-1B	Class-C Amp. (Telegraphy)	3500	500	—	−250	300	40	40	—	30	—	700
													Class-C Amp. (Telephony)	3000	375	—	−200	200	—	55	70000	35	—	400

[1] Discontinued.
[2] Triode connection—screen grid tied to plate.
[3] Dual tube. Values for both sections, in push-pull. Interelectrode capacitances, however, are for each section.
[4] Terminals 3 and 6 must be connected together.
[5] Filament limited to intermittent operation.
[6] Values are for two tubes in push-pull.
[7] Max.-signal value.
[8] Peak grid-to-grid a.f. volts.
[9] Forced-air cooling required.
[10] Average value.
[11] Two tubes triode connected, G_2 to G_1 through 20K Ω. Input to G_2.

TABLE XVIII—KLYSTRONS

Type	Freq. Range-Mc.	Cathode Volts	Cathode Amp.	Base Connec-tions	Typical Operation	Beam Volts	Beam Ma. (Max.)	Beam Watts (Max.)	Control-Electrode Volts	Reflector Volts	Cathode Ma.	R.F. Driving Power Watts [4]	Output Watts
2K25/ 723A-B	8702–9548	6.3	0.44	Fig. 60	Reflex Oscillator	300	32	—	—	−130/−185	25	—	0.033
2K26	6250-7060	6.3	0.50	Fig. 60	Reflex Oscillator	300	25	—	—	−65/−120	—	—	0.120
2K-28 [5]	1200–3750	6.3	0.65	Fig. 61	Reflex Oscillator	300 [7]	45	—	300	−155/−290	30	—	0.140
2K33	23500–24500	6.3	0.65	Fig. 62	Reflex Oscillator	1800 [7]	—	—	−20/−100	−80/−220	[6]	—	0.04
2K34	2730–3330	6.3	1.6	Fig. 58	Oscillator-Buffer *	1900	150	450	−45	—	75	—	10–14
2K35	2730–3330	6.3	1.6	Fig. 58	Cascade Amplifier *	1500	150	450	0	—	75	0.005	5
2K41	2660–3310	6.3	1.3	Fig. 59	Reflex Oscillator *	1000	60	75	+24	−510	60	—	0.75
2K42 [3]	3300–4200	6.3	1.3	Fig. 59	Reflex Oscillator *	1000	60	75	0	−650	45	—	0.75
2K43 [3]	4200–5700	6.3	1.3	Fig. 59	Reflex Oscillator *	1000	60	75	0	−320	40	—	0.8
2K44 [3]	5700–7500	6.3	1.3	Fig. 59	Reflex Oscillator *	1000	60	75	0	−700	43	—	0.9
2K39 [3]	7500–10300	6.3	1.3	Fig. 59	Reflex Oscillator *	1000	60	75	0	−660	30	—	0.46
2K46	2730–3330 [1] 8190–10000 [2]	6.3	1.3	Fig. 58	Frequency Multiplier *	1500	60	60	−90	—	30	0.01/0.07	0.01–0.07
2K47	250–280 [1] 2250–3360 [2]	6.3	1.3	Fig. 58	Frequency Multiplier *	1000	60	60	−35	—	50	3.5	0.15
2K56	3840–4460	6.3	5.0	Fig. 60	Reflex Oscillator	300	25	—	—	−85/−150	—	—	0.090
3K21 [3]	2300–2725	6.3	1.6	Fig. 58	Oscillator-Amplifier *	2000	150	450	0	—	125	1–3	10–20
3K22 [3]	3320–4000	6.3	1.6	Fig. 58	Oscillator-Amplifier *	2000	150	450	0	—	125	1–3	10–20
3K23 [3]	950–1150	6.3	1.6	Fig. 59	Reflex Oscillator *	1000	90	80	0	−300	70	—	1–2
3K27 [3]	750–960	6.3	1.6	Fig. 59	Reflex Oscillator *	1000	90	80	0	−300	70	—	1–2
3K30 (410R) [3]	2700–3300	6.3	1.6	Fig. 58	Oscillator-Amplifier *	2000	150	450	0	—	125	1–3	10–20
707B [5]	1200–3750	6.3	0.65	Fig. 61	Reflex Oscillator	300 [7]	45	—	300	−155/−290	30	—	0.140
QK159	2950–3275	6.3	0.65	Fig. 63	Reflex Oscillator	300	45	—	300	−100/−175	20	—	0.150
Z-668	21900–26100	—	—	—	Reflex Oscillator *	1700	—	15	—	−1700/−2300	—	—	0.02

[1] Input frequency.
[2] Output frequency.
[3] Tuner required.
[4] At max. ratings.
[5] Has demountable tuning cavity.
[6] Cathode current specified on each tube.
[7] G2 and G3 voltage.
* Forced-air cooling required.

TABLE XIX—CAVITY MAGNETRONS

Type	Class	Band or Range Mc.	Heater		Maximum Ratings				Typical Operation					
			Volts	Amps.	Anode KV.	Anode Amps.	Duty Cycle	Input Watts	Anode KV.	Anode Amps.	Field Gauss	Pulse μ Sec.	P.P.S.	Peak Pwr. Output KW.
RK2J22	1	3267–3333	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	2250	1.0	1000	265
RK2J23	1	3071–3100	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	2400	1.0	1000	275
RK2J24	1	3047–3071	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	2400	1.0	1000	275
RK2J25	1	3019–3047	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	2400	1.0	1000	275
RK2J26	1	2992–3019	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	2400	1.0	1000	275
RK2J27	1	2965–2992	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	2400	1.0	1000	275
RK2J28	1	2939–2965	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	2400	1.0	1000	275
RK2J29	1	2914–2939	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	2400	1.0	1000	275
RK2J30	1	2860–2900	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	1900	1.0	1000	285
RK2J31	1	2820–2860	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	1900	1.0	1000	285
RK2J32	1	2780–2820	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	1900	1.0	1000	285
RK2J33	1	2740–2780	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	1900	1.0	1000	285
RK2J34	1	2700–2740	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	1900	1.0	1000	285
RK2J36	1	9003–9168	6.3	1.3	13.5	12.0	.002	200	11.5	10.0	2500	1.0	1000	15.0
RK2J38	1	3249–3263	6.3	1.25	6.0	8.0	.012	200	4.9	3.0	Pkg.	1.0	2000	5.0
RK2J39	1	3267–3333	6.3	1.25	6.0	8.0	.002	200	5.4	5.0	Pkg.	1.0	2000	8.7
RK2J48	1	9310–9320	6.3	1.0	16.0	16.0	.002	230	12.0	12.0	4850	1.0	1000	50.0
RK2J49	1	9000–9160	6.3	1.0	16.0	16.0	.0012	180	12.0	12.0	5400	1.0	1000	58.0
RK2J50	1	8740–8890	6.3	1.0	16.0	16.0	.0012	180	12.0	12.0	5400	1.0	1000	58.0
RK2J54	2	3123–3259	6.3	1.5	14.0	15.0	.002	250	11.6	12.5	1400	1.0	2000	45.0
RK2J55	1	9345–9405	6.3	1.0	16.0	16.0	.001	180	12.8	12.0	Pkg.	1.0	1000	50.0
RK2J56	1	9215–9275	6.3	1.0	16.0	16.0	.001	180	12.8	12.0	Pkg.	1.0	1000	50.0
RK2J58	2	2992–3100	6.3	1.5	22.0	15.0	.002	600	10.5	12.5	1450	1.0	2000	50.0
RK2J61A	2	3000–3100	6.3	1.5	15.0	15.0	.002	250	10.7	12.5	1300	1.0	2000	35.0
RK2J62A	2	2914–3010	6.3	1.5	15.0	15.0	.002	250	10.2	12.5	1300	1.0	2000	35.0
RK2J66	2	2845–2905	6.3	1.5	20.0	25.0	.001	400	18.0	25.0	1700	1.0	1000	150
RK2J67	2	2795–2855	6.3	1.5	20.0	25.0	.001	400	18.0	25.0	1700	1.0	1000	150
RK2J68	2	2745–2805	6.3	1.5	20.0	25.0	.001	400	18.0	25.0	1700	1.0	1000	150
RK2J69	2	2695–2755	6.3	1.5	20.0	25.0	.001	400	18.0	25.0	1700	1.0	1000	150
RK4J31	1	2860–2900	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
RK4J32	1	2820–2860	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
RK4J33	1	2780–2820	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
RK4J34	1	2740–2780	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
RK4J35	1	2700–2740	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
RK4J36	1	3650–3700	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2500	1.0	400	750
RK4J37	1	3600–3650	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2500	1.0	400	750
RK4J38	1	3550–3600	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2500	1.0	400	750
RK4J39	1	3500–3550	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2500	1.0	400	750
RK4J40	1	3450–3500	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2500	1.0	400	750
RK4J41	1	3400–3450	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2500	1.0	400	750
RK4J43	1	2992–3019	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
RK4J44	1	2965–2992	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
RK4J53	1	2793–2813	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
RK4J54	1	6875–6775	12.6	3.75	25.0	35.0	.001	650	17.5	30.0	Pkg.	1.0	1000	200
RK4J55	1	6775–6675	12.6	3.75	25.0	35.0	.001	650	17.5	30.0	Pkg.	1.0	1000	200
RK4J56	1	6675–6575	12.6	3.75	25.0	35.0	.001	650	17.5	30.0	Pkg.	1.0	1000	200
RK4J57	1	6575–6475	12.6	3.75	25.0	35.0	.001	650	17.5	30.0	Pkg.	1.0	1000	200
RK4J58	1	6475–6375	12.6	3.75	25.0	35.0	.001	650	17.5	30.0	Pkg.	1.0	1000	200
RK4J59	1	6375–6275	12.6	3.75	25.0	35.0	.001	650	17.5	30.0	Pkg.	1.0	1000	200
RK725A	1	9345–9405	6.3	1.0	16.0	16.0	.001	180	12.0	12.0	5400	1.0	1000	50.0

[1] Fixed-frequency—Pulsed. [2] Tunable—Pulsed.

Index

Charts and Tables

Formulas

Text

The Catalog Section

☆ ☆ ☆

In the following pages is a catalog-file of products of the principal manufacturers who serve the short-wave field. Appearance in these pages is by invitation—space has been sold only to those dependable firms whose established integrity and whose products have met with the approval of the American Radio Relay League.

INDEX OF ADVERTISERS

★ CATALOG SECTION ★

The Radio Amateur's Handbook

THE NEW *DIRECT READING* HRO-50

Now, National presents a great new HRO receiver after more than three years of designing, development and testing. Retaining all the world-famous, performance-proved HRO features, this superb receiver — the finest National has ever made — now incorporates no less than 14 advanced-design innovations. Exhaustive comparative tests indicate the new HRO-50, by far the most modern and versatile in its field, will set an entirely new standard of performance for communication receivers.

14 ALL NEW FEATURES

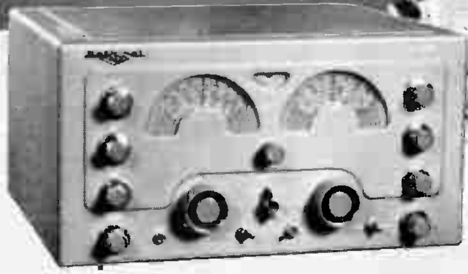
1. Direct frequency reading linear scale with a single range in view at a time. 2. Provisions for using 100/1000 kcs. crystal calibrator unit, switched from panel. 3. Variable front-of-panel antenna trimmer. 4. Built-in power supply with heat resistant barrier. 5. Front-of-panel oscillator compensation control. 6. B.F.O. switch separated from B.F.O. frequency control. 7. Provision for incorporation of NFM adapter inside receiver, switched from front panel. 8. Dimmer control for dial and meter illumination. 9. Miniature tubes in front end and high frequency oscillator. 10. Speaker matching transformer built into receiver with 8 and 500 -ohm output terminals. 11. High frequency and beat frequency oscillator circuits not disabled when receiver in "send" position. 12. High-fidelity push-pull audio amplifier, 8 watts undistorted output. 13. Tip jack for phono input. 14. Accessory socket for Select-o-Ject (see page 4).

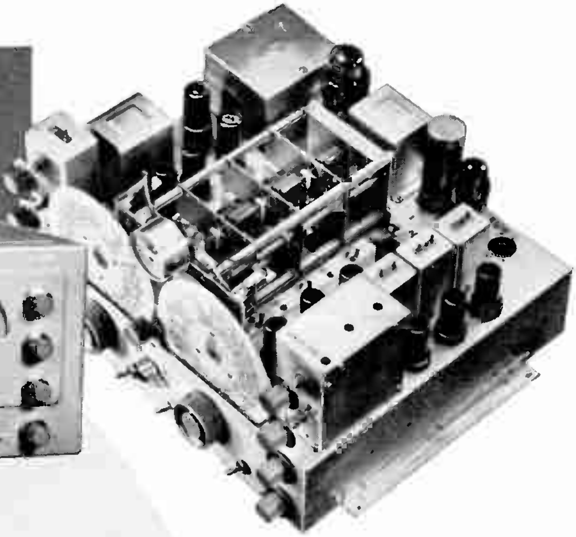
TUBE COMPLEMENT:

1st RF, 6BA6; 2nd RF 6BA6; Mixer, 6BE6; HF oscillator 6C4; voltage regulator OB2; 1st I.F., 6K7; 2nd I.F., 6K7; Det./AVC, 6H6; B.F. Oscillator, 6J7; Noise Limiter, 6H6; 1st Audio, 6SJ7; phase inverter/ "S"-meter amp. 6SN7; Push-pull audio, 2-6V6; Rectifier, 5V4G; accessory crystal calibrator, 6AQ5; NFM adapter I.F. amplifier, 6SK7; Ratio detector, 6H6. Freq. range: 50 kc.-420 kc., 480 kc.-35 mc Coils AA, B, C, and D furnished covering standard amateur 160-10 meter bands.

National
©
EST 1914
NATIONAL COMPANY, Inc.
MALDEN, MASSACHUSETTS

deluxe receiver for optimum reception under all conditions!





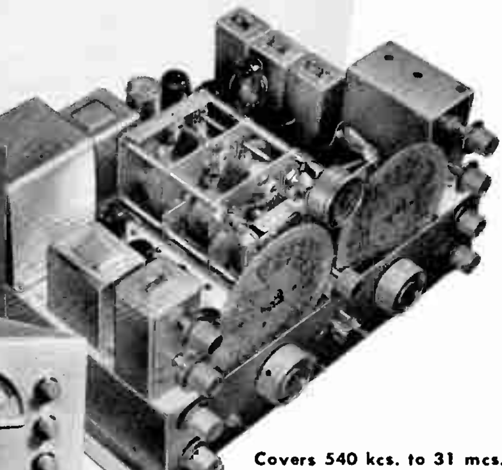
NC-183

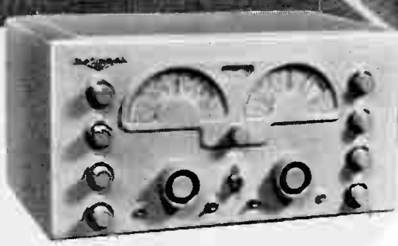
The flawless design and superb construction of this professional communication receiver make possible amazing performance even under the worst operating conditions. If it's possible to receive a signal, the NC-183 will bring it in!

Continuous tuning from 540 kcs to 31 mcs plus the 48 to 56 mcs band for 6-meter reception. Two tuned R.F. stages provide extremely high sensitivity and image rejection. Voltage regulated oscillator and BFO assure minimum drift on phone and CW. Separate main tuning and bandspread dials calibrated for tuning ease. Main dial covers range in five bands. Bandspread dial calibrated for amateur 80, 40, 20, 11-10 and 6-meter bands. Bandspread usable over entire range. Six-position crystal filter provides any selectivity required from very broad to extremely sharp for cutting through adjacent channel interference. New-type noise limiter effectively minimizes electrical interference. High fidelity push-pull audio output with phono input and front-of-panel RADIO-PHONO switch. Accessory socket for NFM adaptor or other unit, such as crystal calibrator. Uses 2-6SG7 R.F.; 16SA7 1st det.; 1-6J5 osc.; 2-6SG7 I. F.; 1-6H6 2nd det.; 1-6SJ7 B.F.O.; 1-6AC7 A.V.C.; 1-6H6 noise limiter; 1-6SJ7 A.F.; 1-6J5 phase inv.; 2-6V6GT aud. out.; 1-VR-150 volt. reg.; 1-5V4G rect. Accessory socket for Select-o-Ject (see page 4).

$268 net*
(Less speaker)

the record-breaking choice of experienced amateurs the world over!





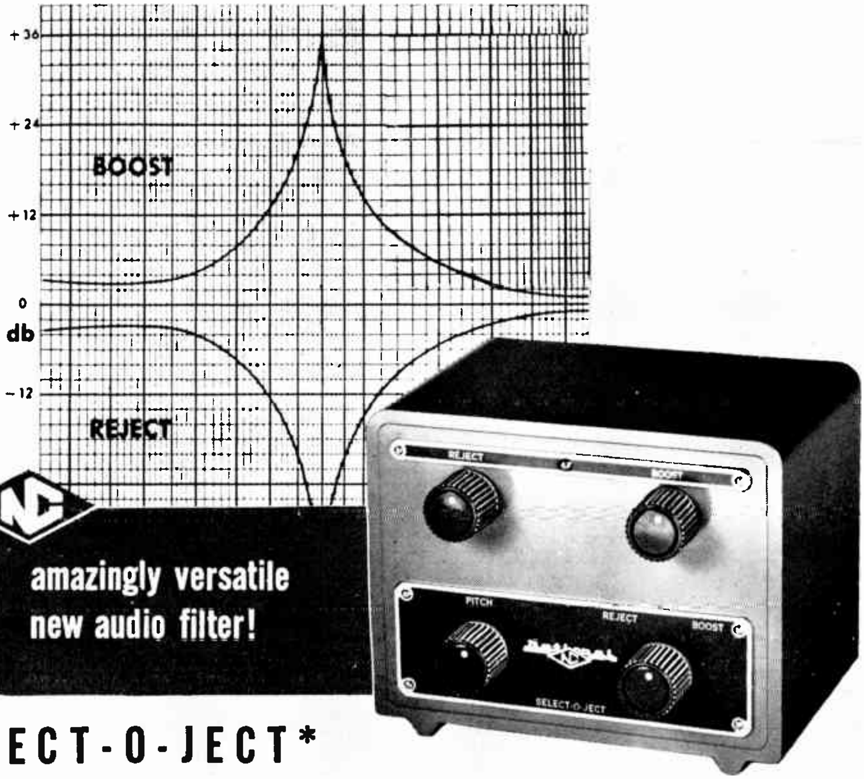
NC-173

The *only* moderate-priced receiver built to National's world-famous standards of sound construction and truly professional performance! Thousands of these sets now in operation attest its popularity and performance.

Covers 540 kcs. to 31 mcs. plus 48 to 56 mcs. for amateur 6-meter band with average sensitivity of 3 microvolts. Separote bandspread dial calibrated for 80, 40, 20, 10 and 6 meter bands. New double-diode noise limiter with variable threshold effective on both phone and CW. Separate AVC usable on phone and CW. New wide-range, 6-position crystal filter, S-meter, antenna trimmer for maximum performance with any antenna, phono input. 1-6SG7 tuned R.F.; 1-6SA7 1st det.; 1-6J5 osc.; 2-6SG7 I.F.; 1-6H6 2nd det. — AVC; 1-6AC7, AVC; 1-6SJ7 BFO; 1-6H6 noise limiter; 1-6SJ7 audio; 1-6v6 output; 1-VR150 volt. reg.; 5Y3GT/G rect.

$199.50 net*

**Slightly higher west of the Rockies.*



amazingly versatile new audio filter!

SELECT-O-JECT*

BOOSTS 38 db! REJECTS 38 db! ANY SELECTED FREQUENCY!

SOJ-1 for all receivers
SOJ-2 wired for HRO-50, NC183 or NC-173
$24.95 net*

* Patent applied for. Manufactured under exclusive agreement with Dr. O. G. Villard, Jr., Engineering Dept., Stanford University.

Set SELECT-O-JECT for REJECT, tune by ear and — presto! — an annoying heterodyne or other unwanted signal practically disappears without materially affecting the wanted signal! Set SELECT-O-JECT for BOOST, tune — and — presto! — a selected signal rises above background noise and interfering signals! Can also be used as audio oscillator having over 100 to 1 frequency range with a single rotation of the tuning knob! Excellent as a code practice oscillator! Effective on any frequency from 80 c.p.s. to 9,000 c.p.s.! This is the amazing circuit described in the November 1949 issue of QST, page 11. See your National dealer for details.

outperforms receivers costing twice as much!

NC-57

Built with all the engineering know-how and craftsmanship of National's more expensive receivers, the NC-57 combines features never found before at tnis low price! The set used by a recent winner of a DX contest sponsored by the internationally famous Shortwave Club of London. Both phone and CW reception over entire frequency spectrum from 550 kcs to 55 mcs in 5 bands. Built-in power supply and PM speaker — nothing else to buy. Voltage stabilized oscillator circuit keeps signal steady regardless of line voltage fluctuations. Automatic threshold noise limiter minimizes interference due to ignition noise, static, etc.

Controls include Main Tuning, Bandspread Tuning, Band Switch, RF Gain, RF Trimmer, BFO-MVC-AVC, ANL Switch, AF Gain, BFO Pitch, Tone Control and On-Off Switch.

Superhet uses: 6SG7 RF amp., 6SB7Y conv., 2-6SG7 IF amp., 6H6 Det., AVC, ANL, 6SN7 Audio amp., BFO, 6V6GT Audio amp., 5Y3GT rect., VR-150 voltage rect. Antenna terminals for single, double or co-ax antenna lead-in. Provision made for connecting external "S" meter plus other accessories. 105-120 V, 50-60 cyc. AC. Gray enamel finish. 16½" x 11¼" x 8¾". Wt. 33 lbs.

$89.50 net*

most popular and versatile VHF design in the field!

HFS

Here is the perfect answer to the need for compact, dependable and versatile VHF reception. Can be used as a complete receiver in itself or as a VHF converter with any receiver tuning to 10.7 mcs. As converter, makes features of connected receiver usable on VHF. Covers entire high frequency spectrum from 27 mcs to 250 mcs — receivesA M, FM and CW with amazing selectivity and sensitivity.

Two-gang Main Tuning Capacitor, panel-controlled Antenna Trimmer Capacitor and 6 sets of plug-in coils tune the receiver in six bands. Power furnished by separate unit. Power supply listed below is excellent where 115-230 V, 50-60 cycle AC is available. Also operates with combination of "B," and storage batteries or 6 volt vibrator-type supply. Wt. 25 lbs.

$142.00* net

Power Supply, 15 lbs., $22.43* net

the ideal receiver for shipboard use or shortwave listening!

NC-57M

Combining versatility, dependability, exceptional sensitivity, and extended frequency range, the NC-57M is ideal as a personal receiver aboard ship or in the shortwave listener's home. Offers continuous frequency range from 540 kcs to 35 mcs plus 200 kcs to 400 kcs. Receives voice, music, and CW code. Bandspread action on any desired frequency assures optimum selectivity. Covers U.S. and European broadcast bands plus shortwave. Scales are marked to show location of such features as amateur, police and foreign frequencies. Voltage regulated oscillator assures excellent stability, regardless of line changes. Built-in power supply for operation from 110/120 volts, either AC or DC. 220-volt operation possible by insertion of external ballast resistor in power plug. Tubes include 6SG7 RF, 6SB7-Y conv.; 6SG7 1st IF; 6SG7 2nd IF; 6H6 2nd det., AVC, ANL; 6SL7 GT/G 1st audio, CWO; 25L6GT aud. out.; OA3/VR-75 volt. reg.; 25Z6GT rect.

$89.50* net

feature for feature — biggest receiver dollar value!

NC-33

Now at last you can get a top-notch communication receiver designed and built by the world-famous National Company at a price that compares favorably with the lowest in the market! Packed with features found in no other receiver at the price!

Four tuning bands provide continuous coverage from 500 KC to 35 MC. Main tuning and bandspread capacities connected in parallel on all bands for bandspread operation at any frequency within tuning range. Amateur, police and foreign broadcast bands clearly identified.

Other big set features include: Automatic Noise Limiter, CW oscillator and pitch control for adjustment of beat note, and Send/Receive Switch. Output to 5" speaker or phone jack which cuts out built-in speaker when headphones are in use. Tunes international SOS frequency. Front-panel mounted controls include: Main tuning, band selector switch, beat oscillator pitch control, code-phone switch, noise limiter switch, and audio gain.

New superhet circuit uses latest type high efficiency tubes. 105-125 V, 50-60 cycles AC or DC.

$57.50* net

**Slightly higher west of the Rockies.*

FOR COMPLETE INFORMATION ON INDIVIDUAL RECEIVERS WRITE

ATIONAL COMPANY INC.

61 SHERMAN ST., MALDEN, MASS.

POPULAR National COMPONENTS

FWG Net $.60
A Victron terminal strip for high frequency use. The binding posts take banana plugs at the top, and grip wires through hole at the bottom, simultaneously, if desired.

FWH Net $.66
The insulators of this terminal assembly are moulded R-39 and have serrated bosses that allow the thinnest panel to be gripped firmly, and yet have ample shoulders. Binding posts same as FWG above.

FWJ Net $.54
This assembly uses the same insulators as the FWH above, but has jacks. When used with the FWF plug (below), there is no exposed metal when the plug is in place.

FWF Net $.70
This molded R-39 plug has two banana plugs on 3/4" centers and fits FWG, FWH or FWJ above. Leads may be brought out through the top or side.

FWA, Post **Net, each $.20**
Brass Nickel Plated

FWE, Jack **Net, each $.15**
Brass Nickel Plated

FWC, Insulator
Net, per pair $.24
R-39 Insulation.

FWB, Insulator
Net, each $.15
Polystyrene insulation.

XS-6 **Net, each $.12**
A low-loss steatite bushing for 1/2" holes. Passes 6-32 screw.

XP-6 **Net, box of ten $.51**
Same as above but polystyrene.

TPB **Net, per dozen $.75**
A threaded polystyrene bushing with removable .093 conductor moulded in, 1/4" diam., 32 thread.

XS-7, (3/8" Hole) **Net $.36**

XS-8, (1/2" Hole) **Net $.48**

XS-1, (1" Hole) **Net $.72**

XS-2, (1 1/2" Hole) **Net $.81**

Prices listed are per pair, including metal fittings and steatite insulators.

XS-9 **Net $.30**
Feed-through insulator. Hole size 13/64". Insulators are adjustable on silver-plated terminal stud for different partition thicknesses. Ceramic insulators are of high grade materials designed for high frequency equipment.

AA-3
A low-loss steatite spreader for 6 inch line spacing. (600 ohms impedance with No. 12 wire.)

AA-5
A low-loss steatite aircraft-type strain insulator.

AA-6
A general purpose strain insulator of low-loss steatite.

GS-1, 1/2" x 1 3/8"
GS-2, 1/2" x 2 7/8"
GS-3, 3/4" x 2 7/8"
GS-4, 3/4" x 4 7/8"
GS-4A, 3/4" x 6 7/8"

Cylindrical low-loss steatite standoff insulators with nickel plated caps and bases.

GSJ, (not illustrated)

A special nickel plated jack top threaded to fit the 3/4" diameter insulators GS-3, GS-4 & GS-4A.

GS-10, 3/4" high

GS-10S (not illustrated) but same as GS-10 except includes threaded stud in top end.

GS-5, 1 1/4" high
GS-6, 2" high
GS-7, 3" high

These cone type standoff insulators are of low loss steatite. They are molded with a tapped hole in each end for mounting as follows:

GS-5, 8-32 tap 7/16" deep; GS-6 & GS-7, 10-24 tap 11/16" deep; GS-10, 6-32 tap 1/4" deep and GS-10S as noted above.

GS-8, with terminal
GS-9, with jack

These low-loss steatite standoff Insulators are also useful as lead-through bushings.

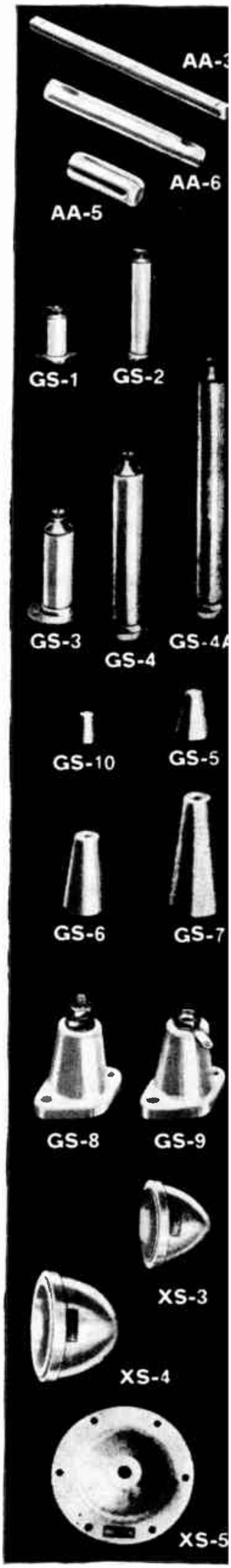
XS-3, (2 3/4" hole)
XS-4, (3 3/4" hole)

Prices are per pair and include nickel plated spindles, lugs and hardware. These low-loss steatite bowls are ideal for lead-in purposes at high voltages.

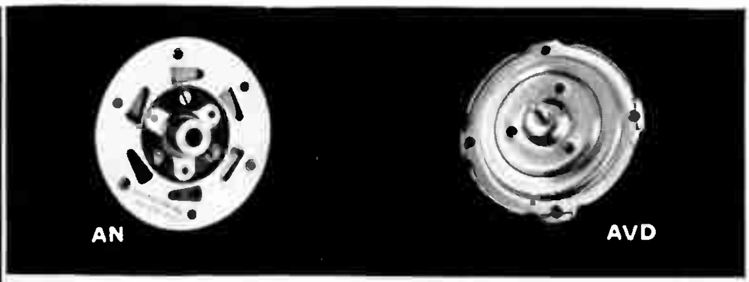
XS-5, Without Fittings

XS-5F, With Fittings

These big low-loss bowls have an extremely long leakage path and a 5 1/4" flange for bolting in place. Insulation steatite. Fittings include nickel plated brass spindles, lugs, nuts and washers.



POPULAR National COMPONENTS



HRT (gray or black) **Net $.75**

The HRT knob is 2$\frac{1}{8}$" in dia. and fits $\frac{1}{4}$" shafts. This knob has a chrome appearance circle and combined with the HRS series shown below gives the new look to panel layouts.

HRS (gray or black) **Net $.50**

The HRS series knobs are a popular easy to grip knob. They are molded of high quality plastic and have 1$\frac{3}{8}$" dia. chrome plated bevel skirts fit $\frac{1}{4}$" shafts available in the following scales:

HRS-1	ON-OFF	through 30°
HRS-2	5-0-5	through 180°
HRS-3	0-10	through 300°
HRS-4		Single etched line

HR (gray or black) **Net $.30**

An HRS type knob without the chrome plated skirt but with a white dot for spotting relative control settings.

HRB **Net $.45**

Ideal for bandswitching or other applications where a switch is turned to several index positions, the new HRB lever knob has just the right feel — a bright zinc alloy die casting.

SB **Net $.18**

A nickel plated brass bushing $\frac{1}{2}$" dia. (Fits $\frac{1}{4}$" shaft).

ODL **Net $.33**

A locking device which clamps the rim of O, K, L and M Dials. Brass, nickel plated.

ODD **Net $.42**

Vernier pinch drive for O, L, or other plain dials.

AN Vernier Mechanism **Net $1.80**

A vernier mechanism ratio 5-1 has an insulated output shaft coupling for $\frac{1}{4}$" shafts. Drive Shaft fits 3/16" knob.

AVD Vernier Mechanism **Net $1.65**

Similar to AN-Output shaft coupling is non insulated.
For commercial uses many variations available. Write for further particulars.

R **Net $ .60**

This small dial has a 1$\frac{5}{8}$" dia. scale calibrated 0-10 in 180° for increased reading with clockwise rotation. Black bakelite knob. Fits $\frac{1}{4}$" shaft.

HRP-P **Net $ .24**

Black bakelite knob 1$\frac{1}{4}$" long and $\frac{1}{2}$" wide. Equipped with pointer. Especially suitable for use on wafer and other rotary switches on laboratory equipment and the like. (Fits $\frac{1}{4}$" shaft).

HRP **Net $ .18**

The type HRP knob has no pointer but is otherwise the same as the knob above. Recommended for uncalibrated or hard-tuning controls. (Fits $\frac{1}{4}$" shaft)

HRK **Net $ .57**

Black bakelite knob 2$\frac{3}{8}$" dial — extremely rugged. This is the knob used on National type O and type L dials.

HRT-M **Net $ .50**

This is a smaller version of the HRT and was designed originally for use on the NC-57 Receiver — now available in choice of gray or black — is 1-7/16" in diameter.

POPULAR National COMPONENTS

N Dial Net $4.50

AD Dial Net $3.00

The four-inch N and AD Dials have engine divided and die stamped scales respectively. The N Dial has a decimal vernier: the AD Dial employs a pointer. The planetary drive has a ratio of 5 to 1, and is contained within the body of the dial. 2, 3, 4, 5 or blank scale. Fits 1/4" shaft. Specify scale.

B Dial Net $2.70

"Velvet Vernier" Dial, Type B, has a compact veriable ratio 6 to 1 min., 20 to 1 max. drive that is smooth and trouble free. The case is black bakelite. 1 or 5 scale. 4" dia. Fits 1/4" shaft. Specify scale.

BM Dial Net $2.10

The BM Dial is a smaller version of the B for use where space is limited. The drive ratio is fixed. Although small in size, the BM Dial has the same smooth action as the larger units. 1 or 5 scale. 3" dia. Fits 1/4" shaft. Specify scale.

AM Dial Net $2.25

The original "Velvet Vernier" mechanism in a metal skirted dial 3" in dia. ratio 5 to 1. It is available with 2, 3, 4, 5 or 6 scale and fits 1/4" shaft.

P Dial Net $1.00

The new P dial is the same as the AM except direct drive.

Type O, 3 1/2" dia., scale 2, with HRK knob, fits 1/4" shafts. Net $1.00

Type L, same as O except 5" dia., scale 2 only. Net $1.95

Type K, same as O except less knob, complete with ODD vernier drive, scale 2 only. Net $1.50

Type M, same as K except 5" dia., scale 2 only. Net $2.25

The dials at the right are for individual calibration: all four employ the noted 5:1 drive ratio Velvet Vernier mechanism and are of excellent quality.

MCN Dial Net $2.70

The MCN dial has been scaled down to lend itself ideally to mobile installations and small converters and tuners. It may also be mounted on the standard 3 1/2" rack panel where such mounting may be desirable. The dial provides three calibrating scales and a 0-100 logging scale. On the rear side of the dial, the mechanism extends 1/4" below the dial frame. 2 3/4" H. x 3 7/8" W.

SCN Dial Net $3.00

The SCN dial provides the same dial scales as the ACN dial but in a reduced size. It is used where economy of panel-mounting space is desirable and where a smaller dial would be out of proportion with the size of the panel. 4-7/16" H. x 6 1/4" W.

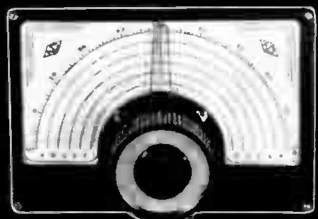
ICN Dial Net $6.00

The ICN dial meets those hundreds of requests from amateurs the world over for an illuminated ACN dial. Two dial lights mounted on the top corners of the dial provide efficient and even illumination on all bands. The dial window has been blanked out in semi-circular shape to prevent shadow casting. Dial scales are the same as those used on the ACN dial. 5 1/8" H. x 7 1/4" W.

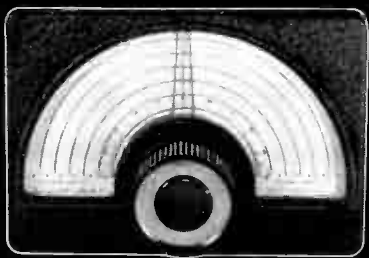
ACN Dial Net $3.30

The ACN is the original of this type dial, a National design for the benefit of experimenters who "build their own" and desire direct calibration. 5" H x 7 1/4" W.

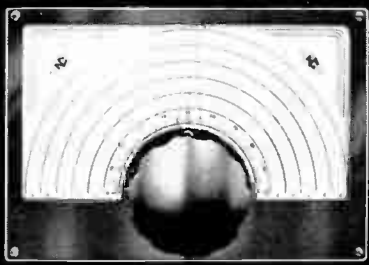
MCN



SCN



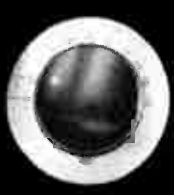
ICN



ACN

DIAL SCALES

Scale	Divisions	Rotation	Direction of Condenser Rotation for increase of dial reading
1	0–100–0	180°	Either
2	0–100	180°	Counter Clockwise
3	100–0	180°	Clockwise
4	150–0	270°	Clockwise
5	200–0	360°	Clockwise
6	0–150	270°	Counter Clockwise



AM

O
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P

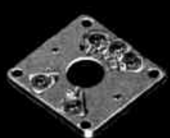
POPULAR National COMPONENTS


XLA


XLA-S


XLA-C


XCA


XMA


XOA-7 (Axial)
XOA-C-7


XOR-7 (Radial)
XOR-C-7


XOA-C-9

XLA **Net $.99**
A low-loss socket for the 6F4 and 950 series acorn tubes for frequencies as high as 600 Mc. Conventional by-pass condensers may be compactly mounted between the contact terminals and the chassis. Low contact resistance, short and direct leads and low and constant inductance are features.

XLA-S **Net $.36**
An internal shield fitting the XLA socket and suitable for tubes such as the 956.

XLA-C **Net $.36**
This miniature by-pass condenser may be mounted inside the socket, directly below the contact. Capacities of 50 or 100 mmf. available.

XCA **Net $.99**
A low-loss steatite socket for acorn triodes. Pin grips are designed to accept tube prongs with minimum strain but exert maximum pressure when seated.

XMA **Net $1.32**
For pentode acorn tubes, this socket has built-in by-pass condensers. The base is a copper plate.

XOA-7 (mica-filled bakelite) **Net $.50**

XOA-C-7 (ceramic) **Net $.50**

XOR-7 (mica-filled bakelite) **Net $.50**

XOR-C-7 (ceramic) **Net $.50**
These high quality sockets for the 7 pin miniature tubes have silver plated beryllium copper contacts that correctly grip the tube pins close to the base of the tube to provide the short leads and low inductance so necessary in ultra-high frequency design.
A novel feature of these new sockets is the interchangeability of the contacts, which are easily removed for replacement. This permits the use of a mixture of axial (XOA) and radial (XOR) type contacts in the same socket to obtain the shortest possible leads, or minimum size in tight places. The above sockets all mount with two 4-40 screws on .875" centers. Chassis cutout should be ¾" dia. Shields for use with these sockets are on page 21.

XOA-C-9 (ceramic) **Net $.57**

XOR-C-9 (ceramic) **Net $.57**
These sockets are for the new 9-pin miniature tubes. The XOR-C-9 (not illustrated) has radial contacts. Both have all of the features described above for the 7-pin types and they also mount with 4-40 screws. Mounting center dimension is 1⅛", the chassis cutout should be 13/16" dia.

CIR SERIES SOCKETS

Any Type **Net $.30**
Always a popular National component, type CIR Sockets feature low-loss steatite insulation, a contact that grips the tube prong for its entire length, and a metal ring for six position mounting.
XC-4, 5, 6, 7S, 7L and CIR-4, 5, 6, 7S and 7L all have 1-27/32" mounting centers. CIR-8E has slotted holes in plate but will mount on 1-27/32" center. CIR-8 and XC-8 have 1½" mounting centers.

XC SERIES SOCKETS

XC-4	**Net $.36**
XC-5	**Net $.39**
XC-6	**Net $.42**
XC-7S	**Net $.45**
XC-7L	**Net $.45**
XC-8	**Net $.39**

National wafer sockets have exceptionally good contacts with high current capacity together with low loss steatite insulation. All types have a locating groove to make tube insertion easy. The XC-6 is ideal for use with AR-17 coils shown on page 24.

HX-29 **Net $.81**
A low-loss wafer socket with steatite insulation for the popular 829 and 832 tubes.

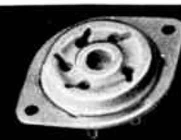
JX-51 **Net $.81**
A low loss steatite wafer socket for the 813 and other tubes having the Giant 7-pin base. (not illustrated)

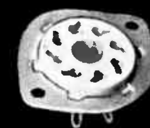
XM-10 **Net $.90**
A heavy duty metal shell socket for tubes having the XU 4-pin base.

XM-50 **Net $1.20**
(see XM-10 for style)
A heavy duty metal shell socket for tubes having the Jumbo 4-pin base ("fifty watters").

HX-100 **Net $.99**
A low loss wafer socket suitable for the type 4-125-A, 4-250-A and other tubes using the Giant 5-pin base. Shield grounding clips are supplied which mount on the chassis with the socket mounting screws to ground the tube shield at three points. Air holes are provided in the socket to permit forced air cooling.

HX-100S **Net $1.65**
Same as above with standoff insulators as illustrated.


CIR-5


CIR-8


CIR-8E


XC-5

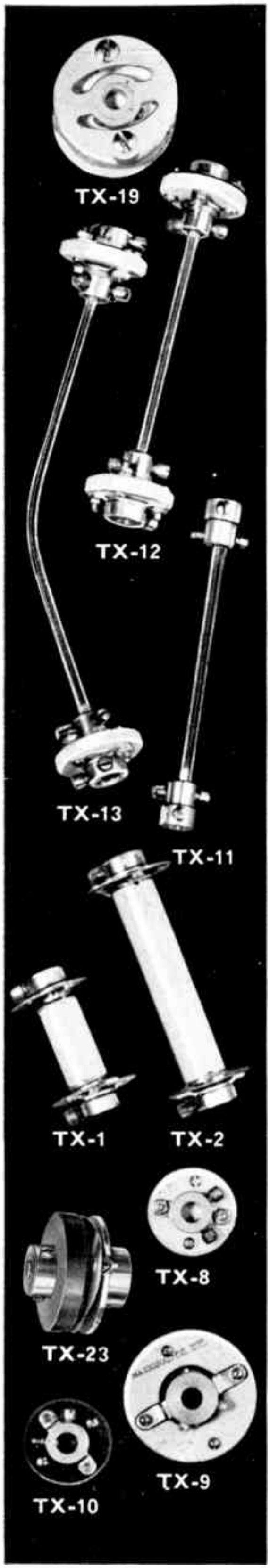

XC-8


HX-29


XM-10


HX-100S

POPULAR National COMPONENTS



SHAFT COUPLINGS

TX-19 **Net $1.25**
A steatite insulated flexible coupling for ¼" shafts. Conservatively rated at 5000 volts peak. Diameter 1⅜", length 1". Length and flashover voltage can be increased by turning callars outboard.

TX-11 **Net $.42**
The flexible shaft of this coupling connects shafts at angles up to 90 degrees, and eliminates misalignment problems. Fits ¼" shafts. Length 4¼".

TX-12, Length 4⅝" **Net $.90**
TX-13, Length 7⅛" **Net $1.05**
These couplings use flexible shafting like the TX-11 above, but are also provided with steatite insulators at each end.

TX-1, Leakage path 1" **Net $.65**
TX-2, Leakage path 2½" **Net $.75**
Flexible couplings with glazed steatite insulation which fit ¼" shafts.

TX-23 **Net $1.35**
A deluxe insulated flexible coupling designed for coupling ¼" shafts. Will handle a maximum radial misalignment of 1/16" also 2 degrees maximum angular misalignment.

TX-24 **Net $1.35**
Same as TX-23, shaft size 5/32".

TX-25 **Net $1.35**
Same as TX-23, non-insulated.

TX-8 **Net $.60**
A non flexible rigid coupling with steatite insulation. 1" diam. Fits ¼" shaft.

TX-10 **Net $.40**
A very compact insulated coupling free from backlash. Insulation is canvas bakelite. 1-1/16" diam. Fits ¼" shaft.

TX-10F (Not illustrated) **Net $.45**
A new version of the TX-10 which employs thin canvas bakelite strips for flexibility.

TX-22 (Not illustrated) **Net $.40**
A non-insulated coupling identical to TX 10 except of all metal construction. Makes good electrical connection between coupled shafts.

TX-9 **Net $.75**
This small insulated flexible coupling provides high electrical efficiency when used to isolate circuits. Insulation is steatite. 1⅝" diam. Fits ¼" shaft.

TX-21 (Not illustrated) **Net $.40**
Similar to TX-10 except 13/16" long and couples ¼" shaft to 5/32" shaft.

SAFETY GRID AND PLATE CAPS

SPP-9 **Net $.21**
Ceramic insulation. Fits 9/16" diameter.

SPP-3 **Net $.21**
Ceramic insulation. Fits ⅜" diameter. National Safety Grid and Plate Caps have a ceramic body which affers protection against accidental contact with high voltage caps on tubes.

GRID AND PLATE GRIPS

Type 12, for 9/16" Caps **Net $.06**
Type 24, for ⅜" Caps **Net $.03**
Type 8, for ¼" Caps **Net $.03**

National Grid and Plate Grips provide a secure and positive contact with the tube cap and yet are released easily by a slight pressure an the ear.

RIGHT ANGLE DRIVES

ACD-1 .... **Net $3.75**
ACD-2 .... **Net $3.90**
ACD-3 .... **Net $3.90**

These sturdy drives were developed for use with the new National AMT condensers (see page 26). They are as compact as the torque requirements will allow and have nickel plated cast frames and bronze gears which operate smoothly without chatter ar binding. The ACD-1 has 32 pitch gears and a ¼" dia. dial shaft and drives ¼" shafts. ACD-2 has 24 pitch gears (for heavier service) and ¼" dia. shaft driving ¼" shafts. ACD-3 is the same as ACD-2 except that it drives ⅜" diameter shafts.

HEAT RADIATING CAPS. Designed to government specifications. Aluminum contact fingers are integral with radiating fins. Tension on fingers maintained by an encircling steel spring. 6 32" tapped center hole for attaching grid ribbon or ather lead. Crimped beryllium copper, silver-plated grid ribbon 3¼" long, supplied with each cap. Special lengths can be supplied to manufacturers in quantities.

Type No.	Price	Hole Size For Lead or Cap	Heat Radiating Connectors To Fit the Following Tubes
HC-26	36¢	.052	3C24-24-24G-25T-27
HC-27	36¢	.062	UH50-HK24-304B-892B-832A-834
HC-28	36¢	.072	35T-35TG-75TH-HK254-HK257B-484-8001
HC-29	50¢	.125	HK57-152TH
HC-30	50¢	.375	4-125A-150TH-2-150D-250R-250TH-250TL-420A-802-803-804-807-808 Grid-814-815-828
HC-31	60¢	.125	304TH-304TL
HC-32	60¢	.570	ZB60-HF60-HF100-111H-211H-203H-HF175-HF300 Grid-100R-HK357C-450TH-454-750TH-805-806-808-809-810-811-812-813-828-833-866-854-1500T-2000T-1054-5331-5332-8000-8003-8005
HC-33	80¢	.810	WL468-WL463-WL460-HF200-HF201-HF300

POPULAR National COMPONENTS

R-100 .... Net $ .35
R-100U .... Net $ .42
R-100S .... Net $ .42
R-100ST .... Net $ .40

These RF chokes are identical electrically, but differ in mounting provisions. The R-100 employs pigtail leads; the R-100U has pigtail leads and a removable stand-off insulator; the R-100S has cotter-pin lug terminals and a non-removable stand-off insulator; the R-100ST has a 6-32 threaded stud at each end. These chokes are available in 2.5, 5 and 10 millihenry sizes and are rated at 125 milliamperes.

R-33 Net $ .35

The R-33 series chokes are 2-section RF chokes available in 10, 50, 100 and 750 microhenry sizes. Also available in this series is a single layer solenoid choke of 1 microhenry inductance. All are rated at 33 milliamperes. The chokes are wound on a ⅝" long form and range in diameter up to 5/16" maximum.

R-50 Net $ .35
R-50-1 Net $ .53

The R-50 series chokes are 3 and 4-section RF chokes and available in 0.5, 1, 2.5, and 10 millihenry sizes. They are rated at 50 milliamperes. The chokes are wound on a 1" long form and have a maximum diameter of 15/32". The 10 millihenry R—50-1 choke is wound on an iron core.

R-33G Net $3.60

The R-33G choke is a 2-section 750 microhenry RF choke hermetically sealed in glass with a current rating of 33 milliamperes. The choke body is 1" long by ⅝" diameter.

R-60 Net $ .35

The R-60 choke is a high current RF choke (500 milliamperes) available in 2 and 4 microhenry sizes. The choke is 1⅛" long by 5/16" diameter.

R-300 .... Net $ .38
R-300U .... Net $ .42
R-300S .... Net $ .42
R-300ST .... Net $ .40

These RF chokes are similar in size to R-100 series but have higher current capacity. The R-300U is provided with a removable stand-off insulator at one end. The R-300S has a non-removable stand-off insulator and cotter-pin lug terminals. The R-300ST has a 6-32 threaded stud at each end. Inductance values of 0.5, 1.0, 2.5 and 5.0 millihenries are available with a current rating of 300 milliamperes. R-300, R-300U, R-300S and R-300ST are identical electrically.

R-152 Net $1.75

For use in the range between 2 and 4 Mc. Ideal for high power transmitter stages operated in the 80 meter amateur band. Inductance 4 m.h., DC resistance 10 ohms, DC current 600 ma. Coils honeycomb wound on steatite core.

R-154 Net $1.75
R-154U Net $1.40

For the 20, 40 and 80 meter bands, Inductance 1 m.h., DC resistance 6 ohms, DC current 600 ma. Coils honeycomb wound on steatite core. The R-154U does not have the third mounting foot and the small insulator, but is otherwise the same as R-154. See illustration.

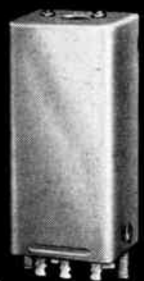
R-175 Net $2.25

The R-175 Choke is suitable for parallel-feed as well as series-feed in transmitters with plate supply up to 3000 volts modulated or 4000 volts unmodulated. Unlike conventional chokes, the reactance of the R-175 is high throughout the 10 and 20 meter bands as well as the 40 and 80 meter bands. Inductance 225 μh, distributed capacity 0.6 mmf., DC resistance 6 ohms, DC current 800 ma., voltage breakdown to base 12,500 volts.

Manufacturers: We have facilities for quantity production of RF chokes of practically any type. Send us your specifications.

POPULAR *National* COMPONENTS

IFC
IFCO



IFL
IFM
IFN
IFO

OSR

XR-50

AR-5

AR-2

I. F. TRANSFORMERS

IFC, Transformer, **Net $4.25**
IFCO, Oscillator, **Net $4.25**
Litz coils wound on a polystyrene form and ceramic insulated air-dielectric trimming condensers make these transformers inherently stable and exceptionally retentive of tuning. The 4½" x 2⅜" x 2" shield can has two 6-32 spade bolts for mounting. Available for either 175 KC or 450-550 KC. Specify frequency.

IFL FM Discriminator **Net $6.90**
IFM IF Transformer **Net $6.45**
IFN IF Transformer **Net $6.45**
IFO FM Ratio Discriminator **Net $6.98**

IFL, IFM, IFN and IFO transformers operate at 10.7 Mc. and are designed for use in FM Superheterodyne receivers. Coils are precision wound on grooved polystyrene forms and tuning is accomplished by movable iron cores. Bandwidth is not affected by tuning slug position. The transformer cans are 1⅜" square and stand 3⅛" above the chassis. Two 6-32 spade bolts are provided for mounting.

The IFL transformer is a 10.7 Mc. FM discriminator transformer suitable for use in conventional FM receiver discriminator circuit and is linear over a band of ±100 Kc.

The IFM transformer is a 10.7 Mc. IF transformer with a 150 Kc. bandwidth at 1.5 db attenuation. Approximate stage gain of 30 is obtained with IFM Transformer and 6SG7 tube.

The IFN transformer is a 10.7 Mc. IF transformer with a 100 Kc. pass band at 1.5 db attenuation. Approximate stage gain of 30 is obtained with IFN Transformer and 6SG7 tube.

The IFO transformer is a 10.7 Mc. FM discriminator transformer of the ratio type and is linear over a band of ±100 Kc.

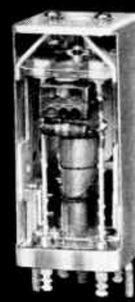
IFJ, with variable coupling **Net $8.25**

IFK, with fixed coupling **Net $7.25**

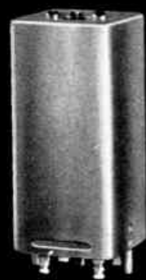
15 Mc. IF transformers suitable for ultra high frequency superheterodynes. They are made in two models with and without variable coupling. Approximate stage gain of 10 is obtained with IFJ or IFK Transformer and 6AB7 tube.

SA:4842 **Net $4.50**
A 456 kc. discriminator transformer for narrow band frequency modulation. This unit is the nucleus of the NFM adapter described by Harrington and Bartell in November 1947 QST. Two slug-tuned secondaries are employed and discrimination is accomplished by resonating one at approximately 10 kc. above, the other at approximately 10 kc. below the center frequency of the i.f. channel.

CD-1, ¼ pint can **Net $.95**
Liquid Polystyrene Cement — is ideal for windings as it will not spoil the properties of the best coil form.



IFJ
IFK



SA-4842

CD-1

COILS AND COIL FORMS

AR-2 High Frequency Coil **Net $1.70**
AR-5 High Frequency Coil **Net $1.46**

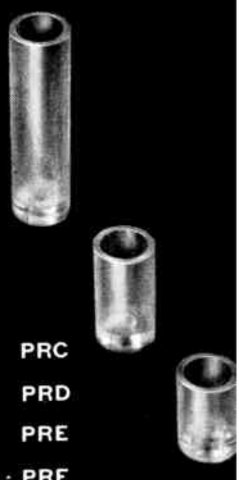
The AR-2 and AR-5 coils are high Q permeability tuned RF coils on low loss mica-filled bakelite forms. The AR-2 coil tunes from 75 Mc. to 220 Mc. with capacities from 100 to 10 mmfd. The AR-5 coil tunes from 37 Mc. to 110 Mc. with capacities from 100 to 10 mmfd. The inductive windings supplied may be replaced by other windings as desired to modify the tuning range.

XR-50 **Net $.90**
These mica-filled bakelite coil forms may be wound as desired to provide a permeability tuned coil. The form winding length is 11/16" and the form winding diameter is ½ inch. The iron slug is ⅜" dia. by ½" long.

OSR **Net $1.80**
A shielded oscillator coil which tunes to 100 kc. with .00041 mfd. Two separate inductances, closely coupled. Excellent for interruption-frequency oscillator in superregenerative receivers.

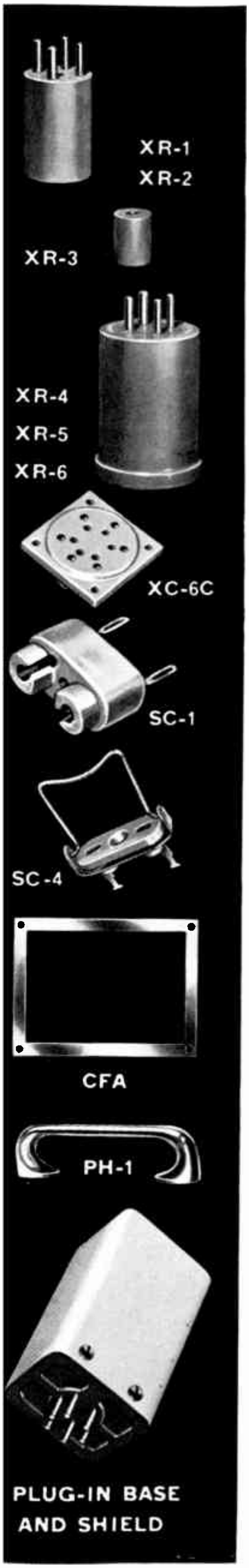
Symbol	Outside Diameter	Length	Net
PRC-1	⅜"	⅜"	.15
PRC-2	⅜"	½"	.15
PRC-3	⅜"	¾"	.15
PRD-1	½"	½"	.15
PRD-2	½"	1"	.15
PRE-1	9/16"	¾"	.18
PRE-2	9/16"	1"	.18
PRE-3	9/16"	2"	.24
PRF-1	¾"	¾"	.24
PRF-2	¾"	1¼"	.30

These small coil forms are of molded polystyrene, open at one end and closed at the other except for a hole which permits mounting by a single 6-32 screw. A size for every application.



PRC
PRD
PRE
PRF

POPULAR National COMPONENTS



Coil Forms molded of R-39 mica-filled bakelite permitting them to be grooved and drilled. Coil Form diameter 1", length 1½".

XR-1, Four Prong — **Net $.35**

XR-2, Without Prongs — **Net .$25**

XR-3, molded of R-39 Diameter 9/16", length ¾" without prongs. **Net $.20**

XR-4, Four Prong — **Net $.51**

XR-5, Five Prong — **Net $.51**

XR-6, Six Prong — **Net $.60**
Molded of R 39 permitting them to be grooved and drilled. Coil Form Diameter 1½" length 2¼". A special socket is required for the XR-6.
National type **XC-6C** — **Net $.51**

SC, Crystal Sockets — **Net $.32**
The SC 1 SC-2 and SC-3 are crystal mounting sockets for crystal holders with mounting pins spaced 0.5000", 0.486", and .750" respectively and pin diameters of 1/8" and 3/32" and 1/8" respectively, steatite insulation. Single 4-36 or 4-40 screw mounting for SC-1 and SC-2; single 6-32 screw mounting for SC-3.

SC-4 Ceramic crystal socket with clamp. Pin spacing .500". Pin dia. 1/32". **Net $.39**

CFA — **Net $.35**
The National chart frame is supplied with a celluloid sheet to cover the chart size 2¼" x 3¼" with sides ¼" wide Durable finish.

PH-1 An attractive and rugged pull handle of cast zinc alloy chrome plated, with 10-32 Tapped Holes on 3¾" mounting centers. **Net $.45**

PH-2 Same as PH-1 but with black or gray finish. **Net $.25**
The plug in base and shield includes the low loss R-39 base which is ideal for mounting condensers and coils when it is desirable to have them shielded and easily removable. Shield is 2" x 2⅜" x 4½".

PB-10-5 — **Net $1.77**
5 Prong base and shield

PB-10-6 — **Net $1.77**
6 Prong base and shield

PB-10-A-5 — **Net $.99**
5 Prong base only

PB-10-A-6 — **Net $.99**
6 Prong base only

RZ Coil Shield — **Net $.35**
1⅜" square x 4" high.
RS Coil Shield — **Net $.35**
1-7/16" x 1⅞" x 3½" high.

RO Coil Shield — **Net $.35**
2" x 2⅜" x 4⅛" high. National Coil Shields are formed from a single piece of pure aluminum. They are mechanically strong and have ample thickness to mount small parts on the walls, and include spade belts, for chassis mounting.

T-78 Tube Shield — **Net $.27**
National Tube Shield type T-78 is a three-piece pure aluminum shield suitable for shielding glass tubes with ST-12 bulb, such as the 6C6 and 6D6 tubes.

JS-1 Jack Shield — **Net $.30**
For shielding small standard jacks mounted behind a panel, or on the ends of extension coils. Indispensable for reducing hum pickup.

XOS Tube Shields — **Net $.48**
The XOS tube shield is a two-piece shield for the miniature Button 7 and 9 pin base tubes. The shield is available in three sizes corresponding to the tube body heights XOS-1 for 1 5/16", XOS 2 for 1½", XOS 3 for 2"

The shield contains a spring which centers tube in shield and holds tube and shield firmly in place.

SHIELDS 7-pin SOCKETS

XOS-1	fit 1 5/16" tube body	**$.48**
XOS-2	fit 1½" tube body	**.48**
XOS-3	fit 2" tube body	**.48**

SHIELDS 9-pin SOCKETS

XOS-4	fit 1 5/16" body	**.51**
XOS-5	fit 1½" tube body	**.51**
XOS-6	fit 2" tube body	**.51**

FXT Fixed tuned exciter tank similar in general construction to National I.F. transformers, this unit has two 25 mmf., 2000 volt air condensers and an unwound XR-2 Coil form.

FXT (Without plug-in base) **Net $3.45**

FXTB-5 (With 5 prong base) **Net $3.90**

FXTB-6 (With 6 prong base) **Net $3.90**

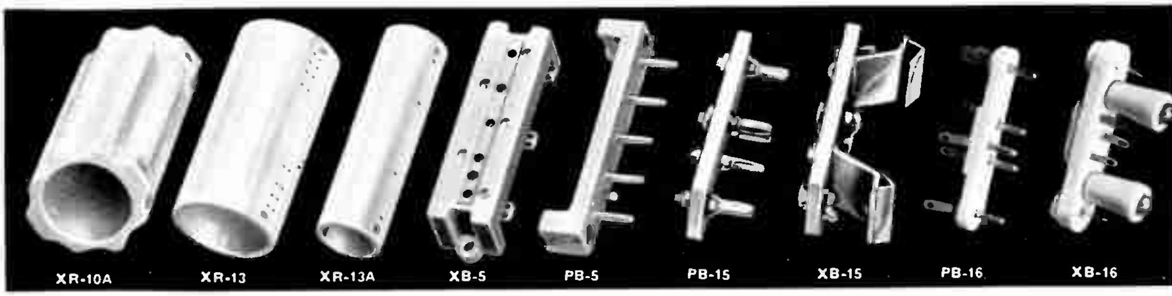
Paint (not illustrated)

CP-1, dark gray — **Net $.40**

CP-2, black — **Net $.40**
A high quality air-drying paint that may be applied with a brush.

CP-3, light gray, matches newest National receivers—for spraying and baking. **Net $.50**

POPULAR National COMPONENTS



TRANSMITTER COIL FORMS

The Transmitter Coil Forms and Mounting are designed as a group, and mount conveniently on the bars of a TMA condenser. The larger coil form, Type XR-14A, (not illustrated) has a winding diameter of 5", a winding length of 3¾" (30 turns total) and is intended for the 80 meter band. The smaller form, Type XR-10A, has a winding length of 3¾" and a winding diameter of 2½" (26 turns total). It is intended for the 20 and 40 meter bands.

Either coil form fits the PB-15 plug. For higher frequencies, the plug may be used with a self-supporting coil of copper tubing. The XB-15 Socket may be mounted on breadboards or chassis, as well as on the TMA Condenser.

SINGLE UNITS

- **XR-10A**, Coil Form only — Net $
- **XR-14A**, Coil Form only — Net $2
- **PB-15**, Plug only — Net $1
- **XB-15**, Socket only — Net $1

ASSEMBLIES

- **UR-10A**, Assembly (including small Coil Form, Plug and Socket) — Net $3
- **UR-14A**, Assembly (including large Coil Form, Plug and Socket) — Net $3

BUFFER COIL FORMS

National Buffer Coil Forms are designed to mount directly on the tie bars of a TMC condenser using the PB-5 Plug and XB-5 Socket. Plug and Socket are of molded R-39.

The two coil forms are of steatite, left unglazed to provide a tooth for coil dope. The larger form, Type XR-13, is 1¾" in diameter and has a winding length of 2¾". The smaller form, Type XR-13A, is 1" in diameter and provides a winding length of 2¾". Both forms have holes for mounting and for leads.

SINGLE UNITS

- **XR-13**, Coil Form only — Net
- **XR-13A**, Coil Form only — Net $
- **PB-5**, Plug only — Net $
- **XB-5**, Socket only — Net $

ASSEMBLIES

- **UR-13A**, Assembly (including small Coil Form, Plug and Socket) — Net $1
- **UR-13**, Assembly (including large Coil Form, Plug and Socket) — Net $1

EXCITER COILS

There is a National exciter coil for every application. AR-15 coils are mounted on 5 pin bases which fit any standard 5 cont tube socket. AR-16 coils are mounted on the well known National PB-16 plug which fits the National XB-16 socket. The AR coils have 6 pin bases which fit standard 6 contact tube sockets and the link windings of this series have center taps which m be grounded for harmonic reduction. All center link models are center tapped for use in balanced circuits. Insulation polystyre and steatite. For use where plate power input does not exceed 50 watts. Available with fixed or swinging end or center links all amateur bands, 6 through 80 meters.

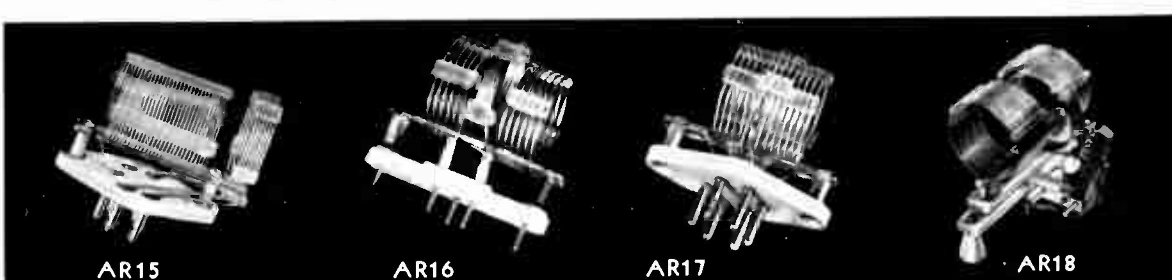
The XR-16 Coil Form (not illustrated) fits the PB-16 Plug-in Base; it has a winding length of 1¾", diameter 1¼"

- **AR-15, AR-16, AR-17** Coil, any type — Net $1.25
- **XR-16** Coil Form — Net $ .42
- **PB-16** Plug-in Base — Net $.
- **XB-16** Socket for PB-16 — Net $

500 WATT COILS

Air-wound coils designed to mount on the split stator models of National AMT condensers. The AR18-C coils have fixed cen links and require the XB18-C socket. The AR18-S coils are designed to accommodate the swinging link furnished with the XB1 socket. Link winding of the XB18-S has a center tap which may be grounded for harmonic reduction. Plugs and jacks are sil plated to insure low contact resistance. Insulation, steatite. The sockets (not illustrated) are 7¼" in length.

AR-18— 6C	$3.25	AR-18—80C	4.50	AR-18—40S	3.95
AR-18—10C	3.50	AR-18— 6S	2.96	AR-18—80S	4.20
AR-18—20C	3.75	AR-18—10S	3.20	XB—18S	4.00
AR-18—40C	4.25	AR-18—20S	3.45	XB—18C	1.50

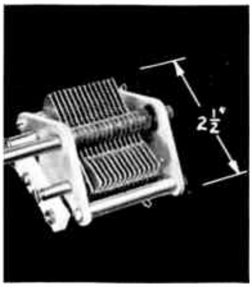


POPULAR National COMPONENTS

EST 1914

TYPE TMS TRANSMITTING CONDENSERS

is a condenser designed for transmitter use in low power stages. It is compact, rigid, and dependable. Provision has n made for mounting either on the panel, on the chassis, or on two stand-off insulators. Insulation is steatite. Voltage rat- listed are conservative.

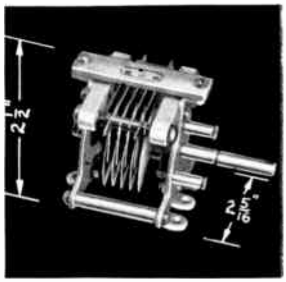


Capacity	Minimum Capacity	Length	Air Gap	Peak Voltage	No. of Plates	Catalog Symbol	Net
SINGLE STATOR MODELS							
100 Mmf.	9.5	3"	.026"	1000v.	9	TMS-100	$2.60
150	11	3"	.026"	1000v.	14	TMS-150	2.80
250	13.5	3"	.026"	1000v.	22	TMS-250	3.30
300	15	3"	.026"	1000v.	27	TMS-300	3.80
35	8	3"	.065"	2000v.	7	TMSA-35	3.90
50	11	3"	.065"	2000v.	11	TMSA-50	4.40
DOUBLE STATOR MODELS							
50–50 Mmf.	6–6	3"	.026"	1000v.	5–5	TMS-50D	$3.00
100–100	7–7	3"	.026"	1000v.	9–9	TMS-100D	3.20
50–50	10.5–10.5	3"	.065"	2000v.	11–11	TMSA-50D	4.40

TYPE TMK TRANSMITTING CONDENSERS

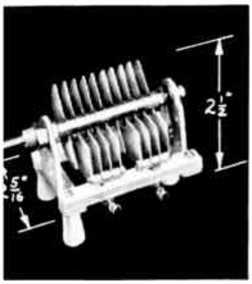
is a new condenser for exciters and low power transmitters. Special provision has been made for mounting AR-16 coils , swivel plug-in mount on either the top or rear of the condenser. For stand-off or panel mounting-steatite insulation.

Capacity	Minimum Capacity	Length	Air Gap	Peak Voltage	No. of Plates	Catalog Symbol	Net
SINGLE STATOR MODELS							
35 Mmf.	7.5	2 7/32"	.047"	1500v.	7	TMK-35	$3.45
50	8	2 3/4"	.047"	1500v.	9	TMK-50	3.55
75	9	2 11/16"	.047"	1500v.	13	TMK-75	3.80
100	10	3"	.047"	1500v.	17	TMK-100	3.95
150	10.5	3 5/8"	.047"	1500v.	25	TMK-150	4.65
200	11	4 1/4"	.047"	1500v.	33	TMK-200	5.25
250	11.5	4 7/8"	.047"	1500v.	41	TMK-250	5.75
DOUBLE STATOR MODELS							
35–35 Mmf.	7.5–7.5	3"	.047"	1500v.	7–7	TMK-35D	$3.80
50–50	8–8	3 5/8"	.047"	1500v.	9–9	TMK-50D	3.95
00–100	10–10	4 1/4"	.047"	1500v.	17–17	TMK-100D	5.25
Swivel Mounting Hardware for AR 16 Coils						SMH	$ .10



TYPE TMH TRANSMITTING CONDENSERS

condenser that features very compact construction. Excellent power factor, and aluminum plates .0400" thick with shed edges. It mounts on the panel or on removable stand-off insulators. Steatite insulators have long leakage path. nd-offs included in listed price.

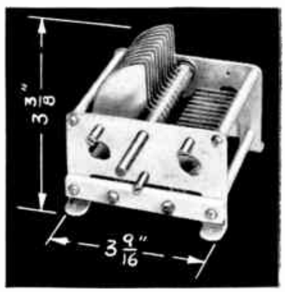


Capacity	Minimum Capacity	Length	Air Gap	Peak Voltage	No. of Plates	Catalog Symbol	Net
SINGLE STATOR MODELS							
50 Mmf.	9	3 3/4"	.085"	3500v.	15	TMH-50	$3.95
75	11	3 3/4"	.085"	3500v.	19	TMH-75	4.15
100	12.5	5 1/8"	.085"	3500v.	25	TMH-100	4.35
150	18	6 1/2"	.085"	3500v.	37	TMH-150	4.95
35	11	5 1/8"	.180"	6500v.	17	TMH-35A	4.25
DOUBLE STATOR MODELS							
35–35 Mmf.	6–6	3 3/4"	.085"	3500v.	9–9	TMH-35D	$4.15
50–50	8–8	5 1/8"	.085"	3500v.	13–13	TMH-50D	4.35
75–75	11–11	6 1/2"	.085"	3500v.	19–19	TMH-75D	4.95

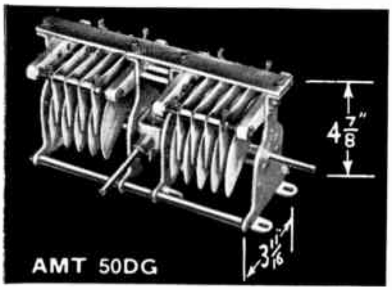
TYPE TMC TRANSMITTING CONDENSERS

condenser designed for use in the power stages of transmitters where peak voltages do not exceed 3000 volts. The frame is emely rigid and arranged for mounting on panel, chassis or stand-off insulators. The plates are aluminum with buffed ies. Insulation is steatite. The stator in the split stator models is supported at both ends.

Capacity	Minimum Capacity	Length	Air Gap	Peak Voltage	No. of Plates	Catalog Symbol	Net
SINGLE STATOR MODELS							
50 Mmf.	10	3"	.077"	3000v.	7	TMC-50	$3.60
100	13	3 1/2"	.077"	3000v.	13	TMC-100	4.25
150	17	4 5/8"	.077"	3000v.	21	TMC-150	5.25
250	23	6"	.077"	3000v.	32	TMC-250	5.70
300	25	6 3/4"	.077"	3000v.	39	TMC-300	6.10
DOUBLE STATOR MODELS							
0–50 Mmf.	9–9	4 5/8"	.077"	3000v.	7–7	TMC-50D	$4.35
00–100	11–11	6 3/4"	.077"	3000v.	13–13	TMC-100D	5.95
00–200	18.5–18.5	9 1/4"	.077"	3000v.	25–25	TMC-200D	7.25



POPULAR National COMPONENTS

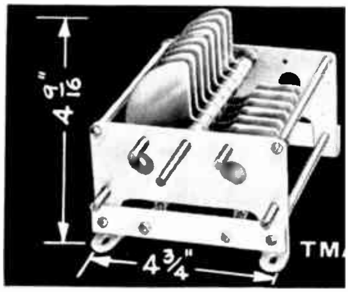


TYPE AMT

A larger and sturdier model of the TMK condenser. The frame is extremely rigid, with mounting feet a part of the end plates. Heavy steatite insulation.

The solid aluminum tie bar across the top of the condenser acts as a mounting for AR-18 series coils in the double stator models.

The double stator models are available in either standard end drive (D series) or center-drive (DG series) with 1/4" dia. shaft extension.



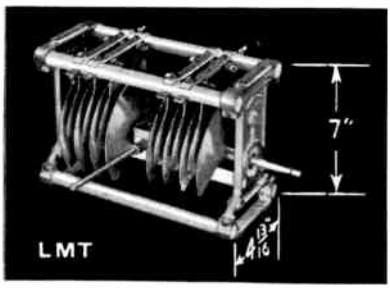
TYPE TMA

This is a larger model of the popular TMC. The frame is extremely rigid and arranged for mounting on panel, chassis or st off insulators. The plates are of heavy aluminum with rounded and buffed edges. Insulation is steatite located outside of concentrated field.

Maximum Capacity	Minimum Capacity	Length	Air Gap	Peak Voltage	No. of Plates	Catalog Symbol	Net
SINGLE STATOR MODELS							
50 Mmf.	13	4 3/4"	.177"	6000 v.	9	AMT-50	$ 5.20
100	20	6 3/4"	.177"	6000 v.	17	AMT-100	6.10
300	19.5	4 9/16"	.077"	3000 v.	23	TMA-300	7.60
50	15	4 9/16"	.171"	6000 v.	7	TMA-50A	4.95
100	19.5	6 7/8"	.171"	6000 v.	15	TMA-100A	5.85
150	22.5	6 7/8"	.171"	6000 v.	21	TMA-150A	6.45
230	33	9 5/16"	.171"	6000 v.	33	TMA-230A	7.95
100	30	9 1/4"	.265"	9000 v.	23	TMA-100B	8.50
150	40.5	12 1/2"	.265"	9000 v.	33	TMA-150B	9.95
50	21	7 1/8"	.359"	12,000 v.	13	TMA-50C	5.55
100	37.5	12 7/8"	.359"	12,000 v.	25	TMA-100C	8.95
75	25	18 1/16"	.719"	20,000 v.	17	TML-75E	18.35
150	60	18 1/16"	.469"	15,000 v.	27	TML-150D	18.50
100	45	13 5/8"	.469"	15,000 v.	19	TML-100D	16.60
50	22	8 5/16"	.469"	15,000 v.	9	TML-50D	11.50
245	54	18 1/16"	.344"	10,000 v.	35	TML-245B	20.15
150	45	13 5/8"	.344"	10,000 v.	21	TML-150B	18.35
100	32	10 15/16"	.344"	10,000 v.	15	TML-100B	17.55
75	23.5	8 5/16"	.344"	10,000 v.	11	TML-75B	12.80
500	55	18 1/16"	.219"	7,500 v.	49	TML-500A	24.60
350	45	13 5/8"	.219"	7,500 v.	33	TML-350A	19.65
250	35	10 15/16"	.219"	7,500 v.	25	TML-250A	18.35
DOUBLE STATOR MODELS D—End drive DG—Center drive							
50–50	13–13	9 3/8"	.177"	6000 v.	18	AMT-50D	7.00
100–100	20–20	13 3/8"	.177"	6000 v.	34	AMT-100D	9.00
50–50	13–13	9 3/8"	.177"	6000 v.	18	AMT-50DG	10.75
100–100	20–20	13 3/8"	.177"	6000 v.	34	AMT-100DG	12.75
200–200	15–15	6 7/8"	.077"	3000 v.	16–16	TMA-200D	9.40
180–180	10–10	12 3/4"	.140"	4000 v.	24–24	TMA-180D	12.90
50–50	12.5–12.5	6 7/8"	.155"	6000 v.	8–8	TMA-50DA	6.75
100–100	17–17	9 5/16"	.155"	6000 v.	14–14	TMA-100DA	8.75
60–60	19.5–19.5	12 1/2"	.249"	9000 v.	15–15	TMA-60DB	8.95
40–40	18–18	12 7/8"	.343"	12,000 v.	11–11	TMA-40DC	8.50
30–30	12–12	18 1/16"	.719"	20,000 v.	7–7	TML-30DE	18.55
60–60	26–26	18 1/16"	.469"	15,000 v.	11–11	TML-60DD	20.15
100–100	27–27	18 1/16"	.344"	10,000 v.	15–15	TML-100DB	22.35
60–60	20–20	13 5/8"	.344"	10,000 v.	9–9	TML-60DB	19.15
200–200	30–30	18 1/16"	.219"	7,500 v.	21–21	TML-200DA	24.60
100–100	17–17	10 15/16"	.219"	7,500 v.	11–11	TML-100DA	20.15

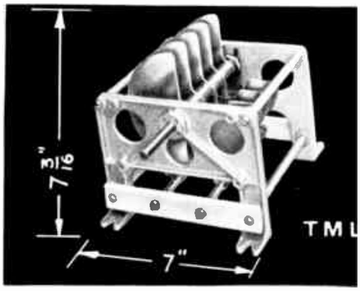
TYPE LMT

A heavy duty transmitting condenser that completely eliminates troublesome closed loops, vastly simplifying the prob of unwanted harmonics. The rotor shaft is completely insulated from the end plates. Long leakage path (higher safety fact Plates and parts are extra heavy with highly polished rounded edges to prevent flash-over. Adjustable stator plate mount and end bearings. Available in single-stator, double-stator, or double-stator right angle center drive models. Same capaci and prices as National TML Condenser. Condensers with right angle drive add $3.90 to price shown.



TYPE TML

is a heavy duty job throughout. The frame structure (rugged aluminum castings with dural tie bars) and precision bearings assure permanent rotor alignment. All plates are extra thick with rounded and polished edges. This, plus specially treated steatite insulators and a husky self-cleaning rotor contact, provides high flashover, current and voltage ratings.



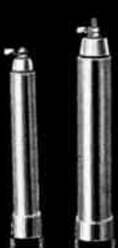
POPULAR COMPONENTS


PSR


M30


W100


NC-600U


TU BY


STN

MINIATURE CONDENSERS:

Type PS variable condensers are compact silver plated units of soldered construction for use as semi-fixed bandsets or padders. Base is steatite — bearing is "snug" but smooth. PSR models are screw-driver adjust type; PSE have ¼" diameter shafts both ends; PSL are similar to PSR but include rotor shaft lock.

Type M-30 **Net $.22**
The M-30 is a tiny (13/16" x 9/16" x ½") mica trimmer — 30 mmf. max. — steatite base.

Type W-75, 75 mmf. **Net $1.60**

Type W-100, 100 mmf. **Net $1.76**
Small air-dielectric padding condensers having a very low temperature coefficient. They are mounted in 1¼" diameter aluminum shields and have ¼" hex heads for socket-wrench adjustment.

The UM condensers are low-loss, aluminum plate staked construction miniature variables designed for UHF converters, VFOs and the like — minimum capacity is exceptionally low. The UMs can be mounted in PB-10 or RO shield cans and have ¼" dia. shafts front and rear for ganging (see pages 21, 23 and 24 for shield cans and couplings). Plates: straight-line-cap., 180° rotation. Dimensions: Base 1" x 2¼", mtg. holes on ⅝" x 1-23/32" centers, 2-5/16" max. length.

The UMB-25 and UMB-50 are differential (balanced stator) models. UM-10D and UMA-25 are double-spaced and the latter is bolted construction for experimental capacity reduction. Hardware for panel or chassis mounting is supplied with all UM condensers.

Capacity	Catalog Symbol			Net
25 mmf.	PSR-25	PSE-25	PSL-25	$1.70
50	PSR-50	PSE-50	PSL-50	1.85
75	PSR-75	PSE-75	PSL-75	2.00
100	PSR-100	PSE-100	PSL-100	2.15

Capacity	Minimum Capacity	No. of Plates	Air Gap	Catalog Symbol	Net
15 mmf.	1.5	6	.017"	UM-15	$1.02
35	2.5	12	.017"	UM-35	1.15
50	3	16	.017"	UM-50	1.25
75	3.5	22	.017"	UM-75	1.45
100	4.5	28	.017"	UM-100	1.60
10	1	8	.042"	UM-10D	1.40
25	3.4	14	.042"	UMA-25	1.75
BALANCED STATOR MODEL					
25	2	4-4-4	.017"	UMB-25	$2.40
50	5	8-8-8	.017"	UMB-50	2.70

NEUTRALIZING CONDENSERS:

NC-600U **Net $.38**
With standoff insulator

NC-600 **Net $.32**
Without insulator

For neutralizing low power beam tubes requiring from .5 to 4 mmf., and 1500 max. total volts such as the 6L6. The NC-600U is supplied with a GS-10 standoff insulator screwed on one end, which may be removed for pigtail mounting.

"TU BY" CONDENSERS

Tubular condensers providing short r f. path between plate and cathode for tubes having the plate connection at the top. Design reduces harmonics and helps eliminate parasitics. 3,000 volts or 1,500 volts. 15 mmfd. **Net $1.80**

STN **Net $2.07**
The Type STN has a maximum capacity of 18 mmf. (3000 V), making it suitable for such tubes as the 809. It is supplied with two standoff insulators.

NC-800A **Net $3.00**
The NC-800A disk-type neutralizing condenser is suitable for the T40, 35TG, 808 and similar tubes. It is equipped with a clamp for locking. The chart below gives capacity and air gap for different settings.

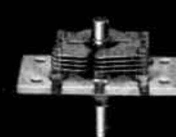
NC-75 **Net $3.60**
For 812, 75TH and similar tubes.

NC-150 **Net $5.25**
For RK36, 100TH, HK354, 250TH, etc.

NC-500 **Net $8.75**
For WE-251, 304TH, 833A and the like. These large disk-type neutralizing condensers are for the higher powered tubes. Disks are aluminum, insulation steatite.


UM


UMA-25


UMB-25


NC-800A


NC-75
NC-150
NC-500

POPULAR National COMPONENTS

PRECISION CONDENSERS

Originally developed for the famous HRO and NC-100 receivers, National PW and NPW condensers and drive units are well known to professional and amateur radio men throughout the world. Sturdily constructed of the finest materials and carefully adjusted by skilled hands, they have become "standard specifications" for applications requiring smooth, precise control and high re-set accuracy.

The Micrometer Dial reads direct to one part in 500. Division lines are approximately 1/4" apart. The drive, at the mid-point of the rotor, is through an enclosed preloaded worm gear with 20 to 1 ratio. Each rotor is individually insulated from the frame, and each has its own individual rotor contact. Stator insulation is steatite. Plate shape is straight-line frequency when the frequency range is 2:1.

PW Condensers are available in 1, 2, 3 or 4 sections, in either 160 or 225 mmf per section. Larger capacities cannot be supplied.

PW-1R	Single section right	**Net $13.50**
PW-1L	Single section left	**Net $13.50**
PW-2R	Double section right	**Net $18.00**
PW-2L	Double section left	**Net $18.00**
PW-2S	Single section each side	**Net $18.00**
PW-3R	Double section right; single left	**Net $24.00**
PW-3L	Double section left; single right	**Net $24.00**
PW-4	Double section each side	**Net $27.00**
NPW-3	Three sections, each 225 mmf.	**Net $24.00**

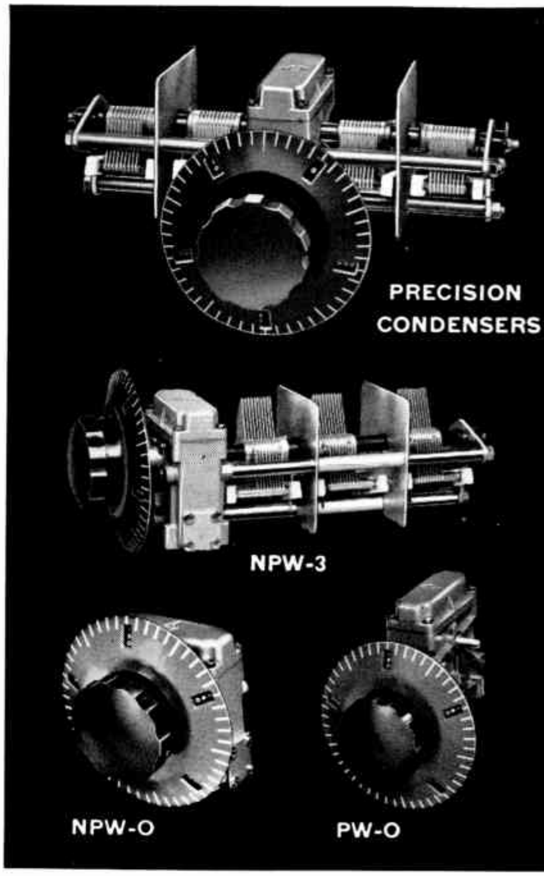
Similar to PW models, except that rotor shaft is perpendicular to panel.

NPW-O **Net $9.00**

Uses parts similar to the NPW condenser. Drive shaft perpendicular to panel. One TX-9 coupling supplied.

PW-O **Net $9.90**

Uses parts similar to the PW condenser. Drive shaft parallel to panel. Two TX-9 couplings supplied.



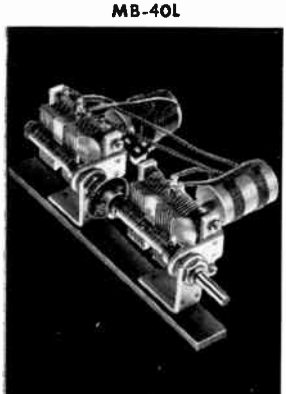
PW-D **Net $5**

The Micrometer Dial used on the condensers and drives above is available separately. It revolves ten times in covering complete range and as there is no gear reduction unit furnished, the driven shaft will revolve ten times, also. The PV dial fits a shaft 5/16" in diameter.

MULTI-BAND TANK ASSEMBLIES

The unique MB-150 Multi-Band Tank tunes all amateur bands from 80 through 10 meters with 180° rotation of the shaft; coils are never changed. The unit is built around a circuit which tunes to two harmonically unrelated frequencies at the sa time. Thus, it becomes possible to cover a wide frequency range and yet maintain a reasonably constant L/C ratio. 3" w x 8 1/4" high (including the GS-10 standoffs) x 9" long overall including the 1/4" dia. shaft and output terminals.

MB-40L



Features of the MB-150:

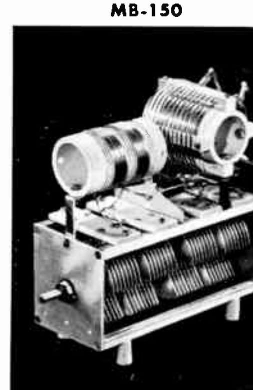
(1) For use as the all-band plate tank in push-pull or single-ended stages running up to 150-watts input (1500 volts peak). It is ideal for a pair of 807s or 809s or a single 829B.

(2) Separate link coupling coil has special clips which adjust to match impedances up to 600 ohms directly. Output couples into a higher powered amplifier, an antenna or an antenna tuning network.

(3) Fast band changing is accomplished without handling coils, thus removing one of the danger points in the amateur station.

MB-150 Multi-Band Tank Assembly **Net $18.75**

MB-150

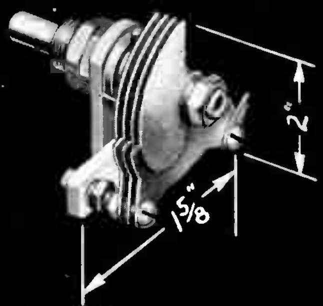


MB 40L LOW-POWER MULTI-BAND TANK

Same principle as the famous MB-150. Logical application as grid circuit for tubes having MB-150 in plate circuit. Will handle 40 watts input if link kept loaded **Net $9.90**

TYPE STHS
STRAIGHT-LINE WAVELENGTH

180° Rotation



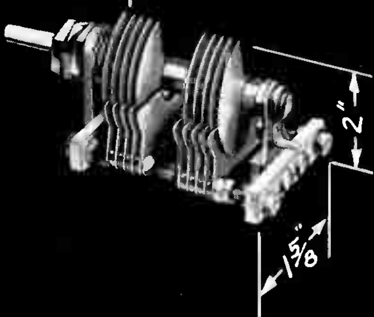
The ST Type condenser has Straight-Line Wavelength plates. All double-bearing models have the front bearing insulated to prevent noise. On special order a shaft extension at each end is available, for ganging. On double-bearing single shaft models, the rotor contact is through a constant impedance pigtail. Steatite insulation.

Capacity	Minimum Capacity	No. of Plates	Air Gap	Length	Catalog Symbol	Net
SINGLE BEARING MODELS						
15 Mmf.	3 Mmf.	3	.018"	1 3/16"	STHS- 15	$1.65
25	3.25	4	.018"	1 3/16"	STHS- 25	1.90
50	3.5	7	.018"	1 3/16"	STHS- 50	2.10

TYPE ST
(Type STD Illustrated)

STRAIGHT-LINE WAVELENGTH

180° Rotation



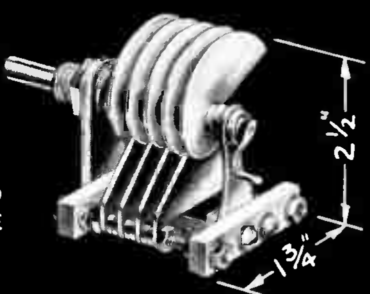
NOTE — Type SS Condensers, having straight-line capacity plates but otherwise similar to the Type ST, are available. Capacities and Prices same as Type ST.

Capacity	Minimum Capacity	No. of Plates	Air Gap	Length	Catalog Symbol	Net
SPLIT STATOR DOUBLE BEARING MODELS						
50–50	5–5	11–11	.026"	2 3/4"	STD- 50	$3.60
100–100	5.5–5.5	14–14	.018"	2 3/4"	STHD-100	3.90
DOUBLE BEARING MODELS						
35 Mmf.	6 Mmf.	8	.026"	2 1/4"	ST- 35	$1.85
50	7	11	.026"	2 1/4"	ST- 50	1.90
75	8	15	.026"	2 1/4"	ST- 75	2.00
100	9	20	.026"	2 1/4"	ST-100	2.10
140	10	27	.026"	2 3/4"	ST-140	2.30
150	10.5	29	.026"	2 3/4"	ST-150	2.30
200	12.0	27	.018"	2 1/4"	STH-200	2.50
250	13.5	32	.018"	2 3/4"	STH-250	2.70
300	15.0	39	.018"	2 3/4"	STH-300	2.90
335	17.0	43	.018"	2 3/4"	STH-335	3.10

TYPE SE
(Type SEU Illustrated)

STRAIGHT-LINE FREQUENCY

270° Rotation

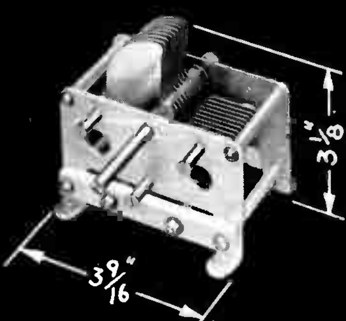


TYPE SE — All models have two rotor bearings, the front bearing being insulated to prevent noise. A shaft extension at each end, for ganging, is available on special order. On models with single shaft extension, the rotor contact is through a constant impedance pigtail. The SEU models (illustrated) are suitable for high voltages as their plates are thick polished aluminum with rounded edges. Other SE condensers do not have polished edges on the plates. Steatite insulation.

15 Mmf.	7 Mmf.	6	.055"	2 1/4"	SEU- 15	$2.80
20	7.5	7	.055"	2 1/4"	SEU- 20	2.95
25	8	9	.055"	2 1/4"	SEU- 25	3.10
50	9	11	.026"	2 1/4"	SE- 50	2.30
75	10	15	.026"	2 1/4"	SE- 75	2.40
100	11.5	20	.026"	2 1/4"	SE-100	2.60
150	13	29	.026"	2 3/4"	SE-150	2.75
200	12	27	.018"	2 1/4"	SEH-200	2.80
250	14	32	.018"	2 3/4"	SEH-250	3.00
300	16	39	.018"	2 3/4"	SEH-300	3.25
335	17	43	.018"	2 3/4"	SEH-335	3.50

TYPE EMC
STRAIGHT-LINE WAVELENGTH

180° Rotation



TYPE EMC — A general purpose condenser available in large sizes and having Straight-Line wavelength plates. They are similar in construction to the TMC Transmitting condenser, and have high efficiency and rugged frames. Insulation is Steatite, and Peak Voltage Rating is 1000 volts. Same sizes available with straight line capacity plates, type DXC condenser.

Capacity	Minimum Capacity	No. of Plates	Length	Catalog Symbol	Net
150 Mmf.	9 Mmf.	9	2 15/16"	EMC-150	$4.50
250	11	15	2 15/16"	EMC-250	4.75
350	12	20	2 15/16"	EMC-350	6.00
500	16	29	4 3/8"	EMC-500	6.75
1000	22	56	6 3/4"	EMC-1000	10.35

McELROY

MANUFACTURING CORPORATION, LITTLETON, MASS.

Telephone, Boston Liberty 2-3411. Cable, Tedmac, Boston

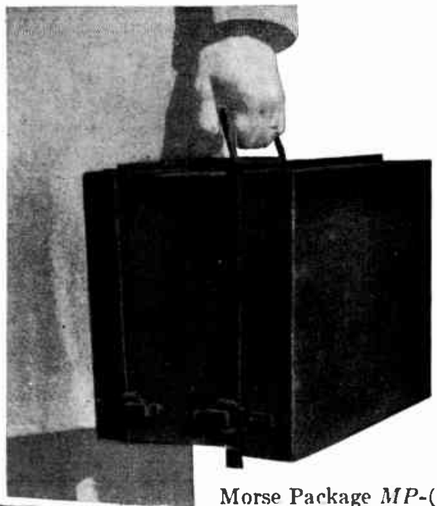
Tens of Thousands of radio operators have been trained on McELROY equipment.

Hundreds of Radiotelegraph Stations — military, naval, commercial — use McELROY equipment.

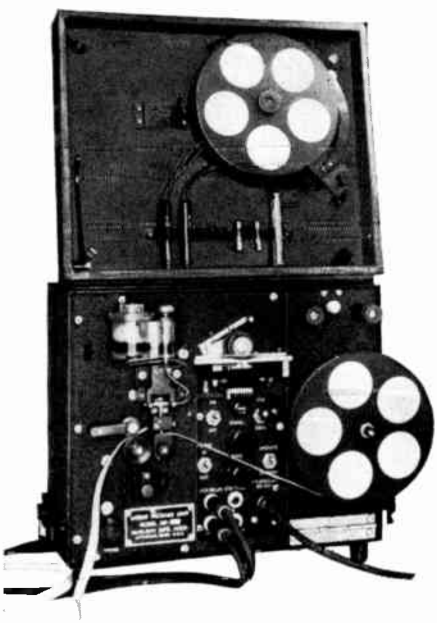
Now even the Smallest Station can convert from manual to automatic operation.

The McElroy Portable Morse Package Unit MP-( )—sturdy, light, compact—gives you perforated tape keyed transmission and inked slip recording at speeds up to 100 words per minute.

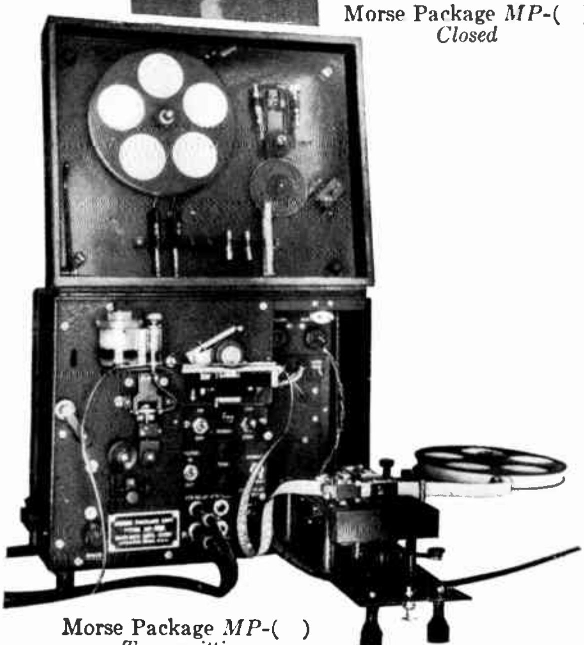
Technical Information will be sent upon request



Morse Package *MP*-( )
Closed



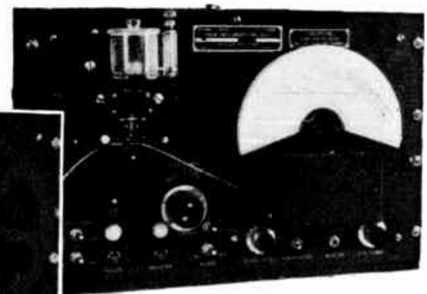
Morse Package *MP*-( )
Receiving



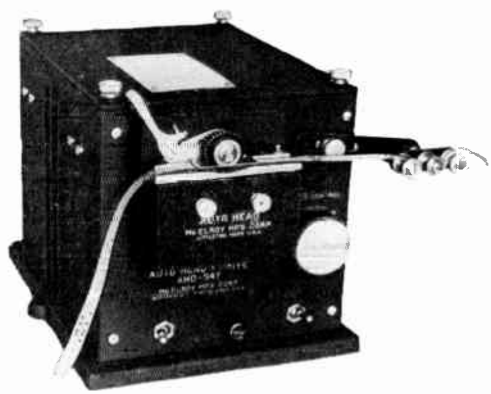
Morse Package *MP*-( )
Transmitting

DEPENDABLE High Speed Communications By SKILLED Communications Engineers of long EXPERIENCE and INTEGRITY

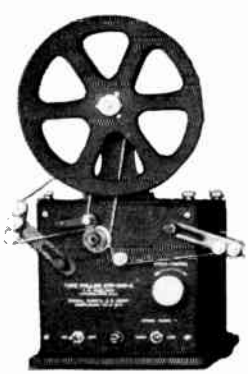
Auto Head, Drive and Keyer Combination *ADK*



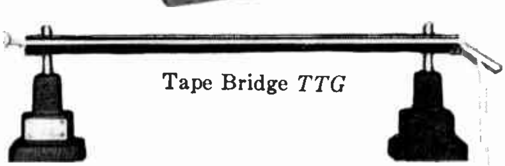
Recorder, Amplifier and Puller Combination *RAPC*



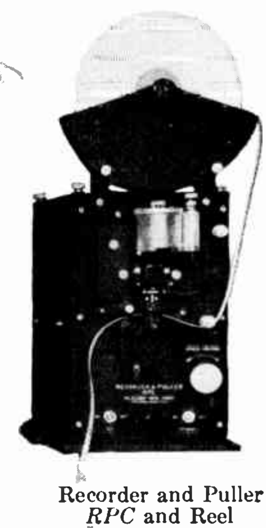
Auto Head and Drive *AHD*



Tape Puller *CTP* and Reel



Tape Bridge *TTG*



Recorder and Puller *RPC* and Reel

Wheatstone Code Perforator *PFR*

Tape Rewinder Model CK-1
Stall-Torque Motor
Teletype or Morse Slip

Outstanding THORDARSON PRODUCTS

The Line Backed by More Than Half a Century of Experience and Leadership!

No matter how specialized or how simple your transformer needs may be, Thordarson can meet them quickly and at remarkably low cost. So vast is Thordarson's experience, that 35,000 different transformers have been designed. All of these active specifications are on file.

The "New Line" utilizes the latest techniques and materials, provides a greater range with few catalog types.

You can depend on Thordarson ratings and on the latest methods of coil impregnation such as:

1. Wax dip or vacuum impregnation.
2. Vacuum varnishing and bake.
3. Fully encased and compound filled.
4. "Tropex" Treated.
5. "Thorseal" — Hermetically sealed.
6. Fully enclosed Hermetically sealed.
7. "Tropex" for complete open frame construction.

AUDIO TRANSFORMERS

Thordarson manufactures a complete line of audio transformers — everything from microphone to speaker. Typical of the standard catalog items are low impedance source (microphone, line or mixer) to single or push-pull grids, transceiver (plate and/or mic to grid), low imp. mic. or v. coil to grid, single plate to single or p.p. grids, p.p. plates to p.p. grids, single or p.p. plates to single or p.p. grids, mic. cable transformers.

POWER EQUIPMENT

Here's truly a famous line, complete from chokes, that include the new swinging-smoothing universal types, power transformers, filament transformers, plate transformers, to vibrator power transformers.

AUDIO BAND REJECTION FILTERS

Thordarson's leadership in this field is the result of units that have exceptionally low insertion losses, sharp attenuation. Four general classifications include: 1. low pass; 2. high pass; 3. band pass; 4. band rejection. Commonly used filters carried in stock. Special ones quoted on application.

AMPLIFIERS

Here, too, Thordarson has a complete line — everything from the phono amplifier to very high power amplifiers for industrial applications.

Illustrated is the new T-32W10. In spite of its modest cost, only $55.00, list, it possesses highest fidelity. It's complete for use with the ordinary high impedance pickup or tuner. Plug-in pre-amplifier, for use with reluctance pick-up or high impedance mike, available at only $9.00, list. Frequency response: 20 to 20,000 cycles. Write for folder.

Industrial Design Sheets

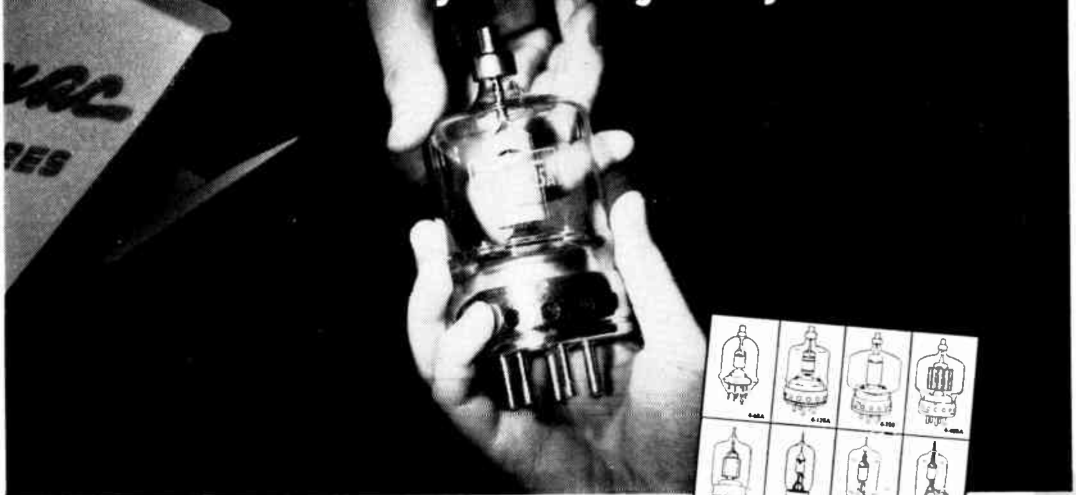
Extremely valuable design data on all phases of transformer and choke characteristics. Available to Industrial groups only. Use company stationery when making request.

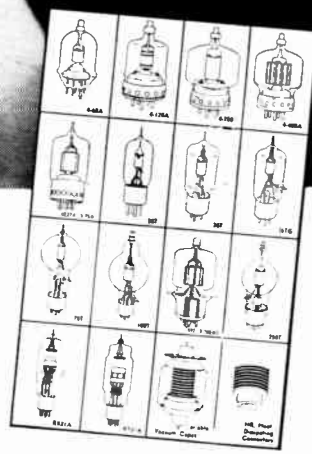
WRITE FOR COMPLETE NEW THORDARSON CATALOG

THORDARSON

ELECTRIC MANUFACTURING DIVISION
MAGUIRE INDUSTRIES, INC.
500 W. Huron St., Chicago 10, Ill.

3 GOOD REASONS WHY . . .
it's to your advantage to buy Eimac Tubes.





Availability

The leading radio parts dealer near you is an Eimac dealer. In his stock are tubes especially suited for every category of amateur service. And, in your locality there is an Eimac factory representative well qualified to assist in your technical problems. Their names, addresses and calls are listed below . . . . drop the one near you a card—he'll be glad to help.

Adequate Data

Comprehensive application data is supplied with each tube. In addition to assist you in designing new or modifying your present equipment, the same data is available without charge by writing . . . Eimac, San Bruno, California. There is also a special packet of data titled "Tubes for Amateur Service." Ask for it.

Proven Performance

Check competition results . . . Eimac tubes are consistently in key-socket position of the transmitters of high scoring stations . . . this is firm evidence of the ability of Eimac tubes to outperform all others.

In commercial electronic applications you also will find Eimac tubes occupying the key socket positions . . . so specified because of top-performance, economy-of-operation, and assured-dependability.

EITEL-McCULLOUGH, INC.
San Bruno, California

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1406 South Grand Ave.
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Eimac TUBES

CONSISTENTLY IN THE WINNERS' CIRCLE

239

hallicrafters

SX-71
$179.50
Amateur Net

S-72
$79.95
Amateur Net

w! A Real Hallicrafters "Ham" Receiver

From the Hams at Hallicrafters to Hams everywhere comes this top-performing receiver in the medium price class. Extra sensitivity, selectivity, and stability, definitely superior image rejection with double superheterodyne circuit, plus built-in Narrow Band FM reception. Surpasses in Ham performance many sets priced much higher.

PERFORMANCE: Continuous coverage 538 kc to 35 Mc and 46—56 Mc. Built-in limiter and balanced detector stages for hiss-free NBFM reception. Double conversion (2075 and 455 kc i-f channels) gives image rejection of better than 150 to 1 at 28 Mc. One r-f, two conversion, and 3 i-f stages yield high gain for sensitivity in the order of 1 microvolt. Sharp selectivity as indicated by the 14 kc band width (1000 times down from resonance) even *before* cutting the crystal filter into the circuit.

CONTROLS: Band Selector 538—1650 kc, 1600—4800 kc, 4.6—13.5 Mc, 12.5—35 Mc, 46—56 Mc. Separate Main and Bandspread tuning; bandspread calibrated for 80, 40, 20, 10, and 6 Meter Bands. BFO Pitch, 3-pos. Selectivity, Xtal Phasing, Tone, AF Gain, RF Gain. ANL, BFO, and Rec. Send switches. "S" Meter adj: on rear.

PHYSICAL DATA: Satin black steel cabinet with chrome trim. Piano hinge top. Size 18½ in. wide, 8⅞ in. high, 12 in. deep. Ship wt. approx. 33 lbs.

EXTERNAL CONNECTIONS: Doublet or single wire antenna. 3.2 and 500-ohm output. Phone jack. Socket for external power. Connections for remote control. 105—125 V 50/60 cycle AC line.

11 TUBES PLUS VOLTAGE REGULATOR AND RECTIFIER: 6BA6 r-f Amp., 6C4 Osc., 6AU6 Mixer, 6BE6 2nd Conv., three 6SK7 i-f Amps., 6H6 ANL and delayed AVC., 6SC7 BFO and a-f Amp., 6AL5 Det., 6K6GT Output, VR-150 Regulator, and 5Y3GT Rectifier.

Matching speaker for SX-71. Two-position tone switch. 500-ohm input. Heavy duty PM type, 6 by 9-inch oval size. 18½ in. wide, 8½ in. high, 9⅝ in. deep. Ship wt. 19 lbs.

R-44B
$24.50

w, All-Wave, AC/DC or Battery Portable

You'll always be in touch with the outside world wherever you go with this new Hallicrafters extra-sensitive portable. Designed both for superior all-wave broadcast reception even in weak signal areas and for Ham operation.

PERFORMANCE: Covers standard broadcast and three short-wave bands 540 kc to 30 Mc continuously. One stage tuned r-f amplification; separate bandspread tuning gang. Two built-in antennas—loop for broadcast and 62-inch telescoping whip for short-wave. Overall sensitivity 1.8 microvolts at 30 Mc, ranging to 6 microvolts at 1.7 Mc.

CONTROLS: Band Selector, r-f Gain, AVC, BFO, a-f Gain, Main tuning, Bandspread tuning.

PHYSICAL DATA: Luggage-type cabinet in brown leatherette. Space inside for phones. Size 14 in. wide, 12¼ in. high, 7¼ in. deep. Ship wt. 16 lbs., less battery pack.

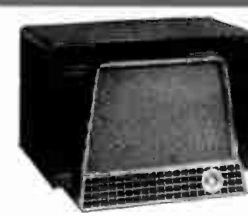
EXTERNAL CONNECTIONS: Phone jack. Antenna terminals if needed. 105—125 V. DC or 50/60 cycle AC line. Battery power 100 ma. at 7.5 V. and 30 ma. at 90 V. Takes RCA VS018, Burgess G6M60, General 60B6F65 and similar packs; life 50 to 100 hours.

8 TUBES PLUS RECT: 1T4 r-f Amp., 1R5 Osc., 1U4 Mixer, two 1U4 i-f Amps., 1U5 Det. and a-f Amp., 1U5 BFO, 3V4 Output, long-life selenium rectifier.

LONG-WAVE MODEL—S-72L. Covers airways radio ranges, control towers, and marine beacons. Range 175—400 kc, .535—12.3 Mc; otherwise identical with S-72......Net $89.95.

hallicrafters

SX-62 .... **$269.50**
Amateur Net



SX-42 .... **$275.00**
Amateur Net

·w "SWL" Version of Famous SX-42

A recent addition to Hallicrafters line and just what the All-Wave listener has been waiting for. Will outperform any ordinary broadcast receiver on any frequency—Standard Broadcast, Short-Wave or FM. Continuous coverage from 540 kc to 109 Mc.

Having basically the same chassis as our best communications receiver, the SX-62 provides communications-receiver performance in simplified form. A single tuning control covers the wide-vision dial. Only one band lights up at a time—you always know just where you are tuning. In addition a 500 kc crystal calibration oscillator is built in, enabling you to adjust the dial pointer to show the exact frequency being tuned at any time.

PERFORMANCE: Continuous AM reception 540 kc to 109 Mc; FM reception 27—109 Mc. Temperature compensated, voltage regulated. Two RF, three IF stages; dual IF channels (455 kc and 10.7 Mc.). Audio flat 50—15,000 cycles; 10-watt push-pull output.

CONTROLS: Band Selector 540—1620 kc, 1.62—4.9 Mc, 4.9—15 Mc, 15—32 Mc, 27—56 Mc, 54—109 Mc. Receive/Standby, Calibration Osc. On/Off, Noise Limiter, Tuning, AF Gain, Phono/FM/AM/CW, six-position Selectivity, four-position Tone, RF Gain, Calibration Reset.

PHYSICAL DATA: Gray steel cabinet with satin chrome trim. Top opens on piano hinge. Cabinet 20 in. wide by 10½ in. high by 16 in. deep. Ship. Wt. 70 lbs.

EXTERNAL CONNECTIONS: Doublet or single wire antenna. 500 and 5000-ohm outputs. Phone jack. Phonograph input jack. Socket for external power. Remote control connections. 105—125 V. 50/60 cycle AC line.

14 TUBES PLUS VOLTAGE REGULATOR AND RECTIFIER: Two 6AG5 RF Amps., 7F8 Conv., 6SK7 IF Amp., 6SG7 IF Amp., 7H7 IF Amp., 7H7 FM Limiter and AM Det., 6H6 FM Det., 7A4 BFO, 6H6 ANL, 6SL7 AF Amp., two 6V6 Push-pull Output, 6C4 Calibration Osc., VR-150 Regulator, 5U4G Rectifier.

R-42 Bass Reflex Speaker $34.50

Matches either SX-62 or SX-42. Two-position tone switch. 500-ohm input. 8-in. heavy-duty PM type. Satin-finish gray metal cabinet 17 in. wide by 11¾ in. high by 12½ in. deep. Ship. wt. 30 lbs.

·ps in Performance and Versatility

Preferred by discriminating Amateurs and SWL's everywhere . . . our best communications receiver! Unsurpassed in versatility and coverage, outstanding in performance. Continuous coverage 540 kc to 110 Mc—Standard Broadcast, Short-Wave and FM.

PERFORMANCE: Continuous AM reception 540 kc to 110 Mc; FM reception 27—110 Mc. Temperature compensated oscillator with voltage regulator. Two RF, three IF stages; dual IF channels (455 kc and 10.7 Mc). Audio flat 50—15,000 cycles; 10-watt push-pull output.

CONTROLS: Band Switch 540—1620 kc, 1620—5000 kc, 5.0—15 Mc, 15—30 Mc, 27—55 Mc, 55—110 Mc. Main tuning dial with logging scale on knob. Band-spread dial calibrated for 3.5, 7, 14, and 28 Mc Bands plus logging scale. Two-position dial lock secures either main or band-spread knobs. AF Volume Control with power switch; AVC, Noise Limiter, Receive/Stand-by switches. Crystal Phasing, AM/FM/CW/Phono, CW Pitch, six-position Selectivity, four-position Tone, RF Gain. "S" Meter adjustment on rear. Control settings for Broadcast and FM Bands marked in color for simplified use by others in family.

PHYSICAL DATA: Gray steel cabinet with satin chrome trim. Top opens on piano hinge. Cabinet 20 in. wide by 10½ in. high by 16 in. deep. Ship. wt. 71 lbs.

EXTERNAL CONNECTIONS: Doublet or single wire antenna. 500 and 5000-ohm outputs. Phone jack. Phonograph input jack. Socket for external power. Remote control connections. 105—125 V. 50/60 cycle AC line.

13 TUBES PLUS VOLTAGE REGULATOR AND RECTIFIER: Two 6AG5 RF Amps., 7F8 Conv., 6SK7 IF Amp., 6SG7 IF Amp., 7H7 IF Amp., 7H7 FM Limiter and AM Det., 6H6 FM Det., 7A4 BFO, 6H6 ANL, 6SL7 AF Amp., two 6V6 Push-pull Output, VR-150 Regulator, 5U4G Rectifier.

hallicrafters

SX-43
$159.50
Amateur Net

S-40A
$79.95
Amateur Net

dest Coverage in Its Class

Here is all you would expect from a truly fine communications receiver plus extra coverage to include the 6-Meter Band and the FM Broadcast Band. Offers coverage, versatility, and performance second only to our SX-42.

PERFORMANCE: AM reception 540 kc to 55 Mc; FM 44—55 Mc and 86—109 Mc. Temperature compensated oscillator. One RF and two IF stages (3rd IF stage above 44 Mc). Dual IF channels (455 kc and 10.7 Mc). Audio response to 10,000 cycles; 3-watt output.

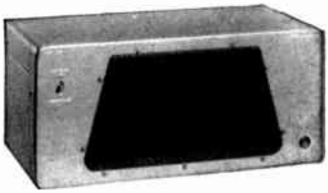
CONTROLS: Band Switch 540—1700 kc, 1700—5000 kc, 5—16 Mc, 14—14.4 Mc, 15.5—44 Mc, 44—55 Mc, 86—109 Mc. Main tuning in Mc; band-spread dial calibrated for 3.5, 7, 14, and 28 Mc bands. Two-position Tone, Receive/Standby and Noise Limiter switches. Crystal phasing, RF Gain, Phono/FM/AM-AVC/AM-MVC/CW, four-position Selectivity, AF Gain, and CW Pitch controls. Adjustment on rear for "S" meter.

PHYSICAL DATA: Gray steel cabinet with satin chrome trim. Piano hinge top. Size 18½ in. wide by 8⅞ in. high by 12 in. deep. Ship. wt. 44 lbs.

EXTERNAL CONNECTIONS: Doublet or single wire antenna. 500 and 5000-ohm outputs. Phono jack. Phonograph input jack. Socket for external power supply. Remote control connections. 105—125 V. 50/60 cycle AC line.

10 TUBES PLUS RECTIFIER: 6BA6 RF Amp., 7F8 Conv., 6SG7 IF Amp., 6SH7 IF Amp., 6SH7 IF Amp., 6H6 AM Det. and ANL, 6AL5 FM Det., 6J5 BFO, 6SQ7 AF Amp., 6V6 Output, 5Y3GT Rectifier.

Matching speaker for SX-43. Two-position tone switch. 500-ohm input. Heavy-duty PM type, 6 by 9-inch oval size. Cabinet size 18½ in. wide by 8½ in. high by 9⅝ in. deep. Ship. Wt. 19 lbs.



R-44
$24.50

nazing Sensitivity and Value

Offers superior performance in the medium price range, born of Hallicrafters long experience in high-quality communications equipment. Complete in itself, with built-in PM speaker.

PERFORMANCE: AM reception 540 kc to 43 Mc. Temperature compensated oscillator. One RF and two IF stages. Audio response to 10,000 cycles.

CONTROLS: Band Switch 540—1700 kc, 1700—5300 kc, 5.3—15.7 Mc, 15.7—43.0 Mc. Main tuning in Mc; band-spread dial has arbitrary scale. AF and RF Gain controls; AVC, BFO, and Noise Limiter switches; three-position Tone, BFO Pitch, and Receive/Standby controls. Settings for Broadcast Band marked in color for simplified use by others in your family.

PHYSICAL DATA: Satin black steel cabinet with brushed chrome trim. Top opens on piano hinge. Size 18½ in. wide by 9 in. high by 9½ in deep. Ship. wt. 32 lbs.

EXTERNAL CONNECTIONS: Doublet or single wire antenna. Phone jack. Socket for external power supply. Remote control connections. 105—125 V. 50/60 cycle AC line.

8 TUBES PLUS RECTIFIER: 6SG7 RF Amp., 6SA7 Conv., two 6SK7 IF Amps., 6H6 ANL and AVC, 6J5GT BFO, 6SQ7, Det. and AF Amp., 6F6G Output, 5Y3GT Rectifier.

hallicrafters

Offers Maximu

S-53 . . . . $69.95

The Radio Tha

S-38A $39.95

erformance in Compact Size

New 2 Mc IF Improves Image Rejection

A recent addition to the Hallicrafters line and a model that is rapidly gaining popularity because of its excellent performance and moderate price. Complete in itself, including built-in PM speaker.

PERFORMANCE: AM reception 540 kc to 31 Mc plus 48—54.5 Mc. Two stages IF with new 2 Mc IF,—high enough to avoid all possible images from ama or stations when operating within the amateur bands.

CONTROLS: Main tuning in Mc; separate band-spread dial with logging scale plus Mc calibration for 48—54.5 Mc band; Receive/Standby switch; Band switch 540—1630 kc, 2.5—6.3 Mc, 6.3—16 Mc, 14—31 Mc, and 48—54.5 Mc; AM/CW; RF Gain, Noise Limiter, AF Gain, two-position Tone; Speaker/Phones switch on rear.

PHYSICAL DATA: Satin black steel cabinet with brushed chrome trim. Top opens on piano hinge. Size 12⅜ in. wide by 7 in. high by 7½ in. deep. Ship wt. 19 lbs.

EXTERNAL CONNECTIONS: Doublet or single wire antenna. Phone tip jacks. Phonograph input jack. 105—125 V. 50/60 cycle AC line.

7 TUBES PLUS RECTIFIER: 6C4 Osc., 6BA6 Mixer, two 6BA6 IF Amps., 6H6 Det., AVC and ANL, 6SC7 BFO and AF Amp., 6K6GT Output, 5Y3GT Rectifier.

maxes Even the Experts

Exceptional Performance at a Low Price

The lowest priced communications receiver on the market . . . with many features found in much higher priced sets. Standard Broadcast plus three Short-Wave bands. Built-in PM speaker.

PERFORMANCE: Continuous AM reception 540 kc to 32 Mc. Maximum sensitivity and selectivity from expertly engineered chassis.

CONTROLS: Main Tuning in Mc; separate band-spread dial with arbitrary scale; Speaker/Phones, AM/CW switches; Band Switch 540—1650 kc, 1.65—5 Mc, 5—14.5 Mc, 13.5—32 Mc; AF Gain, Receive/Standby.

PHYSICAL DATA: Steel cabinet in black wrinkle finish with brushed chrome trim. Size 12⅜ in. wide by 7 in. high by 7½ in. deep. Ship. wt. 14 lbs.

EXTERNAL CONNECTIONS. Doublet or single wire antenna. Phone tip jacks. 105—125 V. DC or 50/60 cycle AC.

4 TUBES PLUS RECTIFIER: 12SA7 Conv., 12SK7 IF Amp. and BFO, 12SQ7 Det. and AVC, 35L6GT Output, 35Z5GT Rectifier.

hallicrafters

The Newest and Most Versatile Transmitter Available

HT-19
$359.50
Amateur Net

Offers Narrow Band FM and CW, plus provisions for AM, to give maximum flexibility on 5 bands. A completely self-contained, medium-power unit for the modern minded amateur. In addition, its compact size and smartly styled cabinet make it especially desirable wherever appearance and space are to be considered.

Consists of an oscillator (crystal controlled or VFO), a frequency modulator with speech amplifier, a buffer and a final amplifier. Extremely high stability and low FM distortion are obtained. The 4-65A in the final, cooled by a 3-inch 800-rpm fan, has a plate input of 185 watts for approximately 125 watts output.

CONTROLS: Operation Switch has three crystal positions plus VFO and NBFM; two pilot lamps show plate and filament power on off; Band Selector switch changes multiplier coils 3.5—4, 7—7.3, 14—14.4, 21—21.45, and 27.16—29.7 Mc; final coils are changed inside the unit, with dummy positions provided for four coils not in use. Check switch turns on oscillator for spotting signals on receiver. Plate switch controls all "B" power and makes connections for remote control. Power switch is in 115-volt line. Deviation Control adjusts for 0.4 ratio on all bands. Osc. Plate Tuning operates osc. gang and calibrated dial. Power Amp. Tuning tunes final plate. Push-button meter switch throws milliammeter from final cathode to final grid.

PHYSICAL DATA: Gray steel cabinet with satin chrome trim. Piano-hinge top with interlock. Size 20 by 10¼ by 16 in. deep. Ship. wt. 98 lbs.

EXTERNAL CONNECTIONS: Microphone connector; keying terminals (osc. keying); 50—600 ohm output (pi-section coupling); six terminals for remote control of either trans. or rec.; four terminals in final screen and plate circuits for applying audio from external modulator for AM. Cord for 105—125 V. 50/60 cycle AC.

5 TUBES PLUS 2 VOLTAGE REGULATORS AND 3 RECTIFIERS: Three 6BA6—Osc., Freq. Modulator, and Speech Amp., 6L6 Buffer, 4-65A Final, VR-150 and VR-105 Regulators, 5Y3GT and two 866 Rectifiers.

Calibrated VFO . . . CW/NBFM

Modernize your present transmitter with this famous Hallicrafters exciter. Crystal or VFO, NBFM or CW on 5 Bands with all coils, speech amplifier, and power supply built in. Features never before available in one low-priced unit. Low frequency drift, low FM distortion, low hum level, excellent keying. Output 2.5 to 4.5 watts. Chassis similar to HT-19 above, less final amplifier.

CONROLS: Operation Switch, Band Selector (ranges like HT-19), Check, Plate, Power, and Deviation switches. Single Tuning control.

PHYSICAL DATA AND CONNECTIONS: Satin gray cabinet, 12⅞ by 7 by 7¾ in. deep. Shipping weight 24 lbs. Microphone, keying, remote control connections. 72-ohm output terminals.

TUBES: Three 6BA6, 6L6, VR-150, VR-105, 5Y3GT.

HT-18
$110.00
Amateur Net

Dependable, All-Purpose Marine Receiver

S-51
$149.50
Amateur Net

PERFORMANCE: Ruggedly constructed for sea or air use with special components to resist salt air, etc. Range 132 kc to 13 Mc. Three pre-tuned channels for fixed-frequency operation. 1020-cycle range filter for better voice reception on ranges. One RF, two IF stages. Temperature compensated.

CONTROLS: RF Gain; Band Selector 132—405 kc, 485—1530 kc, 1450—4550 kc, 4.2—13 Mc plus three fixed frequencies in 200—300 kc and 2—3 Mc ranges; AF Gain, CW/AM, Range Filter, ANL, Tuning, three-position Tone, CW Pitch, Rec./Send.

PHYSICAL DATA AND CONNECTIONS: Gray steel cabinet, 18½ by 9 by 9½ in deep. Ship. wt. 30 lbs. Piano hinge top. Doublet or single wire antenna. Phone jack. Socket for 6, 12, or 32-V. vibrator pack (available separately). 105—125 V. DC or 50/60 cycle AC.

9 TUBES PLUS RECTIFIER: 6SS7 RF Amp., 7A8 Conv., two 6SS7 IF Amp., 7C6 Det., 7A6 ANL, 6SS7 BFO, 35L6GT or 6V6GT Output, 35Z5GT Rect.

Top-quality FM/AM broadcast radio

S-47
$229.50
Amateur Net

Hallicrafters best radio for conventional home reception . . . comparable to chassis found in consoles in the $600—$800 price class. Special features include Automatic Frequency Control, push-button tuning on both Broadcast and FM, and high-fidelity audio.

Automatic Frequency Control gives unprecedented ease of tuning on FM—with unequalled accuracy. As a station is approached, this circuit "takes over" electronically and holds the station in perfect tune with knife-like precision.

PERFORMANCE: Covers Standard Broadcast, FM and three Short-Wave Bands. Two "band-spread" Short-Wave bands spread out stations across the dial for easier tuning of popular foreign broadcasts. Temperature compensated oscillator. One RF, three IF stages. 10-watt push-pull output; audio reponse 30—15,000 cycles for rich, resonant tone.

CONTROLS: Five push buttons for AM and five for FM; Band selector switch FM 88—108 Mc, AM 540—1720 kc, 5.9—18.2 Mc, 9—12 Mc, and 15—18 Mc. Three-position Bass Tone, four-position Treble Tone, Volume, FM tuning, AM tuning.

PHYSICAL DATA: Gray steel, satin chrome trim. Piano hinge top. Size 20 x 10½ x 16 in. deep. Ship. wt. 66 lbs.

EXTERNAL CONNECTIONS: Doublet or single wire antenna. 500-ohm output (speaker not included—use R-42 on Page 3). Phonograph jack; 115 V. outlet for phono motor. Cord for 105—125 V. 50/60 cycle AC.

14 TUBES PLUS RECTIFIER: 6BA6 RF Amp., 6BE6 Mixer, 6J6 HFO and Auto. Freq. Control, two 6SG7 IF Amps., 6SG7 FM 3rd IF Amp. and AM Det., 6SH7 FM 4th IF, 6AL5 FM Det., two 6J5 and two 6SQ7 AF Amps., two 6V6GT Output. 5U4G Rect.

S-47C $209.50
Amateur Net

Chassis Only for Custom Installations

Same FM/AM chassis as in S-47 radio above. Offers superb performance with high-fidelity audio. For custom installations for those who prefer the finest. Size 18½ in. wide by 8¾ in. high by 16 in. deep. Ship. wt. 47 lbs. Fits relay rack.

the hallicrafters company

There is no other radio like a Hallicrafters. Its precision construction and skillful engineering will bring you superb performance on the short wave bands plus fine quality reception of your favorite broadcast programs. Thrilling land, sea, and air communications from all parts of the world plus hours of enjoyment on the amateur bands are yours with a Hallicrafters.

These world famous precision instruments have been sold in 89 different countries, used by 33 governments. They are remembered by veterans, prized by experts, and preferred by radio amateurs who want a radio that is all radio.

hallicrafters TELEVISION

In addition to the high quality communications instruments in this catalog, Hallicrafters also make a complete line of precision-built television receivers—from 7-inch table models to large 19-inch console models. The same advanced engineering that has characterized "the radio man's radio" now brings you improved television with pictures that are exceptionally sharp, bright, and stable.

Complete information on Hallicrafters television models is available in separate folders. Ask wherever Hallicrafters equipment is sold or write direct to—

the hallicrafters co.

4401 West Fifth Ave., Chicago 24, Illinois

All prices in this catalog include Federal Excise Tax, if any. Prices are subject to change without notice.

THERE'S AN RCA TUBE

RCA Power Tube Chart for Amateur Transmitters

CW, FM, AND PHONE TO 30 Mc.

This table of representative tube types has been set up to give suitable choice of tubes for the final and for a preceding stage to drive the final. A choice of buffer, doubler, or oscillator driver stage is provided. The tubes listed have been chosen conservatively to provide ample driving power at 30 Mc. even in circuits having higher than usual losses.

Final Amplifier			Tube Types for Driving Final Amplifier (CW, FM, and Phone)			Class B Modulator
Input Power Watts CW & FM	Phone	Tube Type	As Buffer	As Doubler	As Oscillator	Tube Type
17	15	1-5763	6AK6 6C4 5618 5763	6AK6 6C4 5618 5763	6AK6 6C4 5618 5763	2-6AQ5 (AB) 1-6N7
34	30	2-5763	6C4 5618 5763	6C4 5618 5763	6AK6 6C4 5618 5763	2-6L6 (AB) 2-6F6 (AB)
40	27	1-2E26	2E26 6AG7 6AK6 6F6	2E26 6AG7 6F6 6N7 6V6GT	2E26 6AG7 6F6 6V6GT	2-6L6 (AB_1) 2-6F6 (AB_2)
75	54	1-807	2E26 6AG7 6F6 802 807	2E26 6AG7 6F6 6L6 802 807	2E26 6AG7 6F6 6L6 802 807	2-2E26 2-807 1-815
75	60	1-815				
75	60	2-2E26				
150	120	2-807	2E26 6F6 802 807	2E26 6F6 6L6 6N7 802 807	2E26 6F6 6L6 6V6GT 802 807	2-807 2-811-A
260	175	1-812-A	2E26 802 807	2E26 6L6 802 807	2E26 6L6 802 807	2-807 2-811-A
260	175	1-811-A	2E26 802 807	807 809 811-A 814	807 814	2-807 2-811-A
300	240	1-8005	2E26 802 807	807 809 811-A 814	807 814	2-808 2-811-A 2-8005
345	260	1-4-65A	2E26 6AG7 6AK6 6F6	2E26 6AG7 6F6 6N7 6V6-GT	2E26 6AG7 6F6 6V6-GT	2-811-A 2-8005
500	375	1-4-125A 4D21	2E26 802 807	2E26 6L6 802 807	2E26 6L6 802 807	4-807 2-811-A 2-8005
500	400	1-813				
520	350	2-812-A	2E26 802 807 809 812-A 815	807 809 811-A 814 815	807 814 815	2-808 2-811-A 2-8005
520	350	2-811-A	2-2E26 2-802 807 809 812-A 815	2-807 809 811-A 814	2-807 814 828	2-808 2-811-A 2-8005
600	400	2-808	2-2E26 807 809 811-A 812-A 815	2-807 809 811-A 812-A 814	2-807 814	2-808 2-811-A 2-8005
600	480	2-8005				
690	520	2-4-65A	2E26 802 807	2E26 6L6 802 807	2E26 6L6 802 807	2-811-A 2-8005
750	500	1-8000	807 809 811-A 812-A 814	807 809 811-A 814	807 814	2-811-A 2-8005
750	500	1-810	809 811-A 812-A 814	808 811-A 814 828	not recommended	2-811-A 2-8005
1000	675	1-4-250A 5D22	2E26 802 807 815	2E26 2-6L6 802 807 815	2E26 2-6L6 802 807 815	2-810 2-8000 4-8005
1000	750	2-4-125A 4D21				
1000	800	2-813				
1000	1000	1-833-A	808 809 811-A 812-A 814 8005	808 811-A 814 828	not recommended	2-810 2-8000 4-8005
1000	1000	2-8000	2-807 2-809 811-A 812-A 814	2-809 808 811-A 814	not recommended	2-810 2-8000 4-8005
1000	1000	2-810	2-809 808 811-A 812-A 814 8005	808 811-A 813 828	not recommended	2-810 2-8000 4-8005

FOR EVERY AMATEUR SERVICE

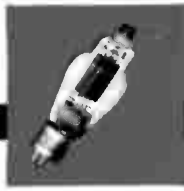
Ratings and Characteristics—Amateur Transmitting Types

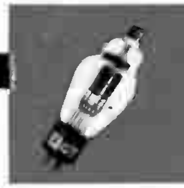
RCA Type	Filament or Heater (H) Volts	Filament or Heater (H) Amps.	Amplification Factor	Max. Frequency for Full Input Mc.	Max. ICAS□ Ratings (Class C Telegraphy) Screen Input Watts	Max. ICAS□ Ratings (Class C Telegraphy) Plate Input Watts
RCA POWER TRIODES						
810	10.0	4.5	36	30	——	750
811-A	6.3	4	160	30	——	260
812-A	6.3	4	29	30	——	260
826	7.5	4	31	250	——	130
832-A	6.3 / 12.6 (H)	1.6	6.5•	200	5*	36*
833-A	10.0	10	35	30	——	1500
8005	10.0	3.25	20	60	——	300
8025-A	6.3	1.92	18	500	——	50
RCA BEAM POWER TUBES						
2E26	6.3 (H)	0.8	6.5•	125	2.5	40
807	6.3 (H)	0.9	8•	60	3.5	75
813	10.0	5	8.5•	30	22	500
815	6.3 / 12.6 (H)	1.6 / 0.8	6.5•	125	4.5*	75*
829-B	6.3 / 12.6 (H)	2.25 / 1.125	9•	200	7*	120*
5763	6.0 (H)	0.75	16•	175	2	17
RCA TETRODES AND PENTODES						
4-65A	6.0	3.5	5•	50	10	450
4-125A/4D21	5.0	6.5	6.2•	120	20	675
4-250A 5D22	5.0	14.5	5.1•	75	35	1400
802	6.3 (H)	0.9	7.3•	30	6	33
5618	3.0	0.46	5.4•	100	2	7.5
RCA RECTIFIERS AND THYRATRONS						
2D21	6.3 (H)	0.6	Gas thyratron, miniature type. Two tubes in grid-controlled, full-wave circuit, up to 80 watts at 400 volts.			
5R4-GY	5.0	2	Full-wave, vacuum rectifier, with choke input, 175 ma. at 750 volts.			
816	2.5	2	Half-wave, mercury-vapor rectifier. Two tubes in full-wave with choke input, 250 ma. up to 2380 volts.			
866-A	2.5	5	Half-wave, mercury-vapor rectifier. Two tubes in full-wave with choke input, 500 ma. up to 3180 volts.			
2050	6.3 (H)	0.6	Gas thyratron. Two tubes in grid-controlled, full-wave circuit, up to 80 watts at 400 volts.			
5557	2.5	5	Mercury-vapor thyratron. Two tubes in grid-controlled, full-wave circuit, up to 1500 watts at 1500 volts.			

RCA GLOW-DISCHARGE (Cold Cathode) TUBES

	Opr. Volts	Current Range	
OA2	151 Volts	5 to 30 ma.	Voltage-regulator types for regulating voltages to oscillators (ECO or XTAL types), oscillator power supplies, to stabilize bias voltages, and for spark-over protection. OA2 and OB2 are miniature types.
OB2	108 Volts	5 to 30 ma.	
OA3	75 Volts	5 to 40 ma.	
OC3	108 Volts	5 to 40 ma.	
OD3	153 Volts	5 to 40 ma.	
5651	87 Volts	1.5 to 3.5 ma.	Voltage reference type for use with series-type of stabilized voltage supply.

□ICAS—Intermittent Commercial and Amateur Service
°Values shown are for Continuous Commercial Service
•Control grid-screen grid mu-factor
*Total for tube





● RCA has a popular tube for every Amateur service, every power, and every active band. A few of the best-known types in each classification are listed.

In addition, there are special-application types, such as *phototubes*, *acorns*, *kinescopes*, *iconoscopes*, and the well-known receiving *types* in metal, glass, and miniature.

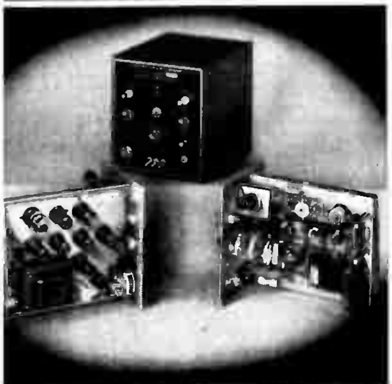
For additional technical data on these RCA tube types, see your local RCA Tube Supplier, or write RCA Commercial Engineering, Section 35AM, Harrison, N. J.

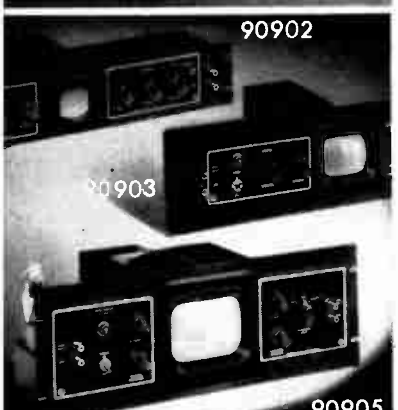
Are you getting RCA HAM TIPS? There's a copy waiting for you at your RCA Tube Supplier.

RADIO CORPORATION of AMERICA

ELECTRON TUBES — **HARRISON, N. J.**

JAMES MILLEN
MALDEN · MASSACHUSETTS





SECONDARY FREQUENCY STANDARD

A precision frequency standard for both laboratory and production uses, adjustable output, provided at intervals of 10, 25, 100 and 1000 kc, with magnitude useful to 50 mc. Harmonic amplifier with tuned plate circuit and panel range switch. 800 cycle modulator with panel control switch. In addition to oscillators, multivibrators, modulators and amplifiers, a built-in detector with phone jack and gain control is incorporated. Self-contained power supply.

Model 90505, with tubes $155.00

ABSORPTION WAVEMETERS

The 90600 series of absorption wavemeters are available in several styles and many different ranges. Most popular is kit of four units, covering range of 3.0 to 140 mc.

Model 90600 .... $18.00

GRID DIP METER

The No. 90651 MILLEN GRID DIP METER is compact and completely self contained. The AC power supply is of the "transformer" type. The drum dial has seven calibrated uniform length scales from 1.5 MC to 300 MC with generous over laps plus an arbitrary scale for use with special application inductors. Internal terminal strip permits battery operation for antenna measurement.

No. 90651, with tube $55.00

LABORATORY SYNCHROSCOPES

The 5" laboratory synchroscopes are available with and without detector-video strips.

Model P-4-2, with tubes $350.00
Model P-4E-2, with tubes 445.00

MINIATURE SYNCHROSCOPE

The compact design of the No. 90952, measuring only 7½" x 5⅝" x 13", and weighing only 17 lbs., makes available for the first time a truly DESIGNED FOR APPLICATION "field service" Synchroscope.

No. 90952, with tubes $375.00

CATHODE RAY OSCILLOSCOPES

The No. 90902, No. 90903 and No. 90905 Rack Panel Oscilloscopes, for two, three and five inch tubes, respectively, are inexpensive basic units comprising power supply, brilliancy and centering controls, safety features, magnetic shielding, switches, etc. As a transmitter monitor, no additional equipment or accessories are required. The well-known trapezoidal monitoring patterns are secured by feeding modulated carrier voltage from a pickup loop directly to vertical plates of the cathode ray tube and audio modulating voltage to horizontal plates. By the addition of such units as sweeps, pulse generators, amplifiers, servo sweeps, etc., all of which can be conveniently and neatly constructed on companion rack panels, the original basic 'scope unit may be expanded to serve any conceivable industrial or laboratory application.

No. 90902, less tubes .... $ 42.50
No. 90903, less tubes 49.50
No. 90905, less tubes 100.00

'SCOPE AMPLIFIER—SWEEP UNIT

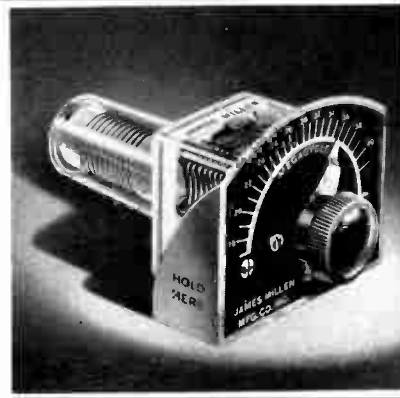
Vertical and horizontal amplifiers along with hard-tube, saw tooth sweep generator. Complete with power supply mounted on a standard 5¼" rack panel.

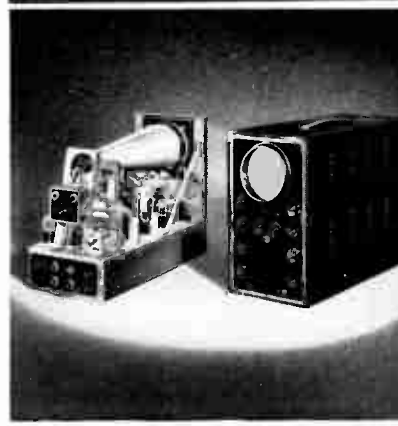
No. 90921, with tubes $75.00

REGULATED POWER SUPPLIES

A compact, uncased, regulated power supply, either for table use in the laboratory or for incorporation as an integral part of larger equipments. 50 watts, with regulated voltage from 0 to 200 volts.

Model 90201, less tubes $100.00





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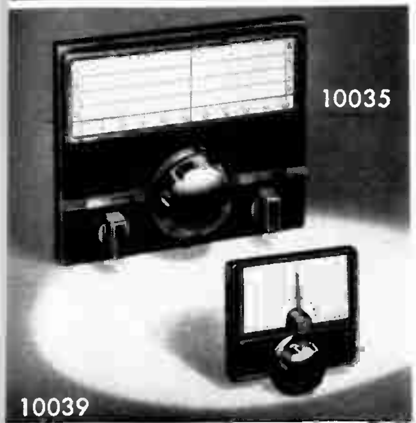
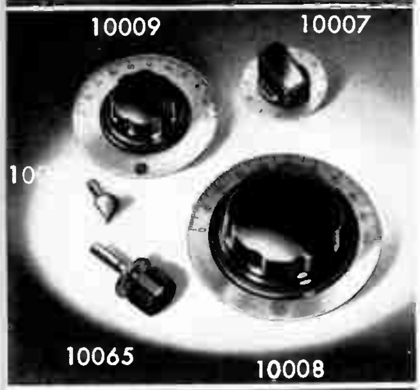




INSTRUMENT DIALS

The No. 10030 is an extremely sturdy instrument type indicator. Control shaft has 1 to 1 ratio. Veeder type counter is direct reading in 99 revolutions and vernier scale permits readings to 1 part in 100 of a single revolution. Has built-in dial lock and ¼" drive shaft coupling. May be used with multi-revolution transmitter controls, etc., or through gear reduction mechanism for control of fractional revolution capacitors, etc., in receivers or laboratory instruments.

The No. 10035 illuminated panel dial has 12 to 1 ratio; size, 8½" x 6½". Small No. 10039 has 8 to 1 ratio; size, 4" x 3¼". Both are of compact mechanical design, easy to mount and have totally self-contained mechanism, thus eliminating back of panel interference. Provision for mounting and marking auxiliary controls, such as switches, potentiometers, etc., provided on the No. 10035. Standard finish, either size, flat black art metal.

No. 10039	$ 2.70
No. 10035	6.00
No. 10030	25.00

DIALS AND KNOBS

Just a few of the many stock types of small dials and knobs are illustrated herewith. 10007 is 1⅜" diameter, 10009 is 2½" and 10008 is 3½".

No. 10007	$ .60
No. 10008	1.00
No. 10009	.85
No. 10021	.15
No. 10065	.45

PANEL MARKING TRANSFERS

The panel marking transfers have ⅛" block letters. Special solution furnished. Must not be used with water. Equally satisfactory on smooth or wrinkle finished panels or chassis. Ample supply of every popular word or marking required for amateur or commercial equipment.

No. 59001, white letters	$1.25
No. 59002, black letters	1.25

HIGH FREQUENCY TRANSMITTER

The No. 90810 crystal control transmitter provides 75 watt output (higher output may be obtained by the use of forced cooling) on the 20, 10–11, 6 and 2 meter amateur bands. Provisions are made for quick band shift by means of the new 48000 series high frequency plug-in coils.

No. 90810, less tubes and crystals	$69.75

HIGH FREQUENCY RF AMPLIFIER

A physically small unit capable of a power output of 70 to 85 watts on 'phone or 87 to 110 watts on C-W on 20, 15, 11, 10, 6 or 2 meter amateur bands. Provision is made for quick band shift by means of the new No. 48000 series VHF plug-in coils. The No. 90811 unit uses either an 829-B or 3E29.

No. 90811 with 10 meter band coils, less tube	$33.00

HIGH VOLTAGE POWER SUPPLY

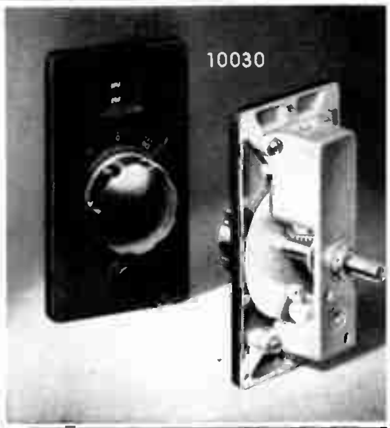
The No. 90281 high voltage power supply has a d.c. output of 700 volts, with maximum current of 250 ma. In addition, a.c. filament power of 6.3 volts at 4 amperes is also available so that this power supply is an ideal unit for use with transmitters, such as the Millen No. 90800, as well as general laboratory purposes. The power supply uses two No. 816 rectifiers and has a two section pi filter with 10 henry General Electric chokes and a 2-2-10 mfd. bank of 1000 volt General Electric Pyranol capacitors. The panel is standard 8¾" x 19" rack mounting.

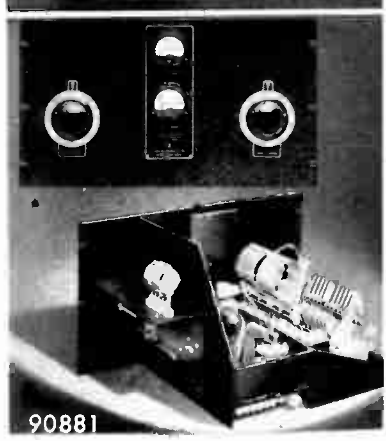
No. 90281, less tubes	$84.50

RF POWER AMPLIFIER

This 500 watt amplifier may be used as the basis of a high power amateur transmitter or as a means for increasing the power output of an existing transmitter. As shipped from the factory, the No. 90881 RF power amplifier is wired for use with the popular RCA or G.E. "812" type tubes, but adequate instructions are furnished for readjusting for operation with such other popular amateur style transmitting tubes as Taylor TZ40, Eimac 35T, etc. The amplifier is of unusually sturdy mechanical construction, on a 10½" relay rack panel. Plug-in inductors are furnished for operation on 10, 20, 40 or 80 meter amateur bands. The standard Millen No. 90800 exciter unit is an ideal driver for the new No. 90881 RF power amplifier.

No. 90881, with one set of coils, but less tubes	$89.50





JAMES MILLEN
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2101


92105

R9'er MATCHING PREAMPLIFIER

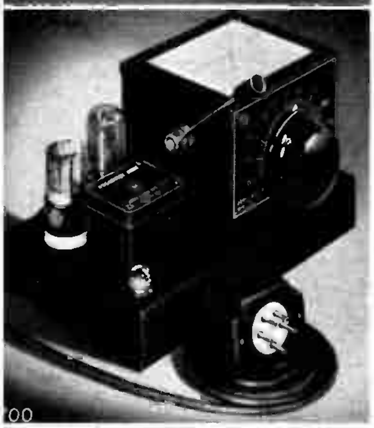
The Millen 92101 is an electronic impedance matching device and a broad band preamplifier combined into a single unit, designed primarily for operation on 6 and 10 meters. Coils for 20 meter band also available.

No. 92101, less tubes $24.75

SINGLE SIDEBAND SELECTOR

The No. 92105 is designed to permit Single Sideband Selection with existing receivers. Full technical details in April 1948 QST. Produced in co-operation and under exclusive U. S. patent license (2,364,863 and others) with the J. L. A. McLaughlin Research Laboratories.

No. 92105 with tubes and crystals $75.00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90711

FREQUENCY SHIFTER

A favorite frequency shifter, plugs in, in place of crystal, for instant finger-tip control of carrier frequency. Low drift, chirpless keying, vibration immune, big band spread, accurate calibration.

Model 90700, with tubes $42.50

VARIABLE FREQUENCY OSCILLATOR

The No. 90711 is a complete transmitter control unit with 6SK7 temperature-compensated, electron coupled oscillator of exceptional stability and low drift, a 6SK7 broad-band buffer or frequency doubler, a 6A67 tuned amplifier which tracks with the oscillator tuning, and a regulated power supply. Output sufficient to drive an 807 is available on 160, 80 and 40 meters and reduced output is available on 20 meters. Close frequency setting is obtained by means of the vernier control arm at the right of the dial. Since the output is isolated from the oscillator by two stages, zero frequency shift occurs when the output load is varied from open circuit to short circuit. The entire unit is unusually solidly built so that no frequency shift occurs due to vibration. The keying is clean and free from all annoying chirp, quick drift, jump, and similar difficulties often encountered in keying variable frequency oscillators.

No. 90711, with tubes $89.75


00

74400

50 WATT TRANSMITTER

Based on an original Handbook design, this flexible unit is ideal for either low power amateur band transmitter use or as an exciter for high power PA stages.

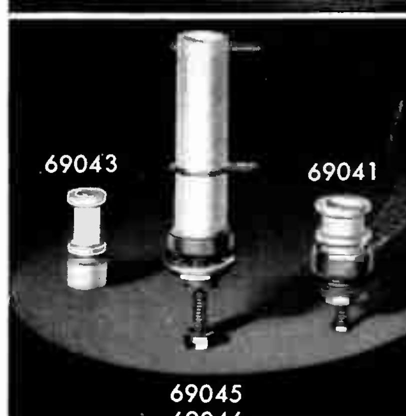
Model 90800, less tubes $42.50

OCTAL BASE AND SHIELD

Low loss phenolic base with octal socket plug and aluminum shield can 1 7/16 x 1 7/8 x 3 15/16.

No. 74400 . . . . $.75


37412
102
FULL SIZE


69043
69041
69045
69046

TRANSMISSION LINE PLUG

An inexpensive, compact, and efficient polyethylene unit for use with the 300 ohm ribbon type polyethylene transmission lines. Fits into standard Millen No. 33102 (crystal) socket. Pin spacing 1/2", diameter .095".

No. 37412 . . . . $.21

PERMEABILITY TUNED CERAMIC FORMS

In addition to the popular shielded plug-in permeability tuned forms, 74000 series, the 69040 series of ceramic permeability tuned unshielded forms are available as standard stock items. Winding diameters and lengths of winding space are 13/32 x 7/32 for 69041–2; 1/4 x 3/8 for 69043–7–8, 1/2 x 1 1/16 for 69045–6; 3/16 x 3/16 for 69044.

No. 69041—(Copper Slug) . . . . $.75
No. 69042—(Iron Core) 75
No. 69043—(Iron Core) 75
No. 69044—(Copper Slug) 75
No. 69045—(Copper Slug) 90
No. 69046—(Iron Core) 90
No. 69047—(Copper Slug) 90
No. 69048—(Iron Core) 90

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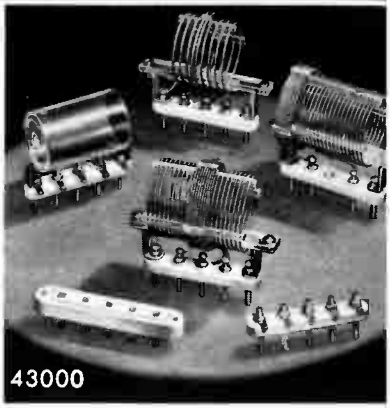


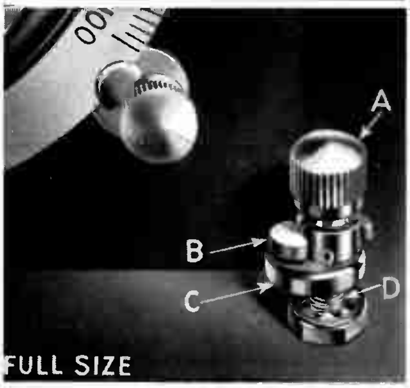
SHAFT LOCKS

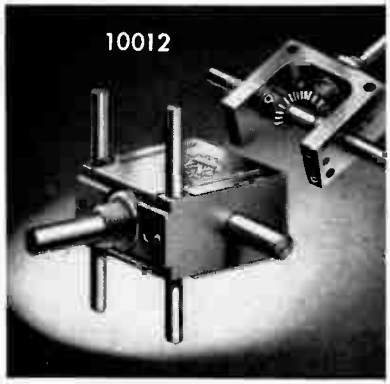
In addition to the original No. 10060 and No. 10061 "DESIGNED FOR APPLICATION" shaft locks, we can also furnish such variations as the No. 10062 and No. 10063 for easy thumb operation as illustrated above. The No. 10061 instantly converts any plain "¼ shaft" volume control, condenser, etc. from "plain" to "shaft locked" type. Each to mount in place of regular mounting nut.

No. 10060	$.36
No. 10061	.36
No. 10062	.45
No. 10063	.45

TRANSMITTING TANK COILS

A full line—all popular wattages for all bands. Send for special catalog.





DIAL LOCK

Compact, easy to mount, positive in action, does not alter dial setting in operation! Rotation of knob "A" depresses finger "B" and "C" without imparting any rotary motion to Dial. Single hole mounted.

No. 10050 .... $.45

RIGHT ANGLE DRIVE

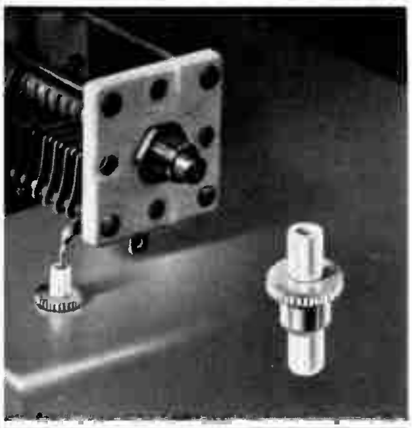
Extremely compact, with provisions for many methods of mounting. Ideal for operating potentiometers, switches, etc., that must be located, for short leads, in remote parts of chassis.

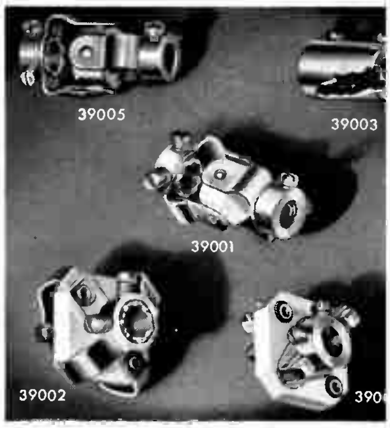
No. 10012 .... $3.75

THRU-BUSHING

Efficient, compact, easy to use and neat appearing. Fits ¼" hole in chassis. Held in place with a drop of solder or a "nick" from a crimping tool.

No. 32150 .... $.05





FLEXIBLE COUPLINGS

The No. 39000 series of Millen "Designed for Application" flexible coupling units include, in addition to improved versions of the conventional types, also such exclusive original designs as the No. 39001 insulated universal joint and the No. 39006 "slide-action" coupling (in both steatite and bakelite insulation).

The No. 39006 "slide-action" coupling permits longitudinal shaft motion, eccentric shaft motion and out-of-line operation, as well as angular drive without backlash.

The No. 39005 is similar to the No. 39001, but is not insulated and is designed for applications where relatively high torque is required. The steatite insulated No. 39001 has a special anti-backlash pivot and socket grip feature. All of the above illustrated units are for ¼" shaft and are standard production type units.

No. 39001	$.42
No. 39002	.42
No. 39003	.21
No. 39005	.42
No. 39006	.42

CATHODE RAY TUBE SHIELDS

For many years we have specialized in the design and manufacture of magnetic metal shields of nicolai and mumetal for cathode ray tubes in our own complete equipment, as well as for applications of all other principal complete equipment manufacturers. Stock types as well as special designs to customers' specifications promptly available.

No. 80045—Nicolai for 5" tube	$10.50
No. 80043—Nicolai for 3" tube	6.00
No. 80042—Nicolai for 2" tube	5.25

BEZELS FOR CATHODE RAY TUBES

Five inch bezel is of cast aluminum with black wrinkle finish. Complete with neoprene cushion, green lucite filter scale and four screws for quick detachment from panel when inserting tube.

No. 80075—5"	$7.50
No. 80073—3"	3.90
No. 80072—2"	1.25

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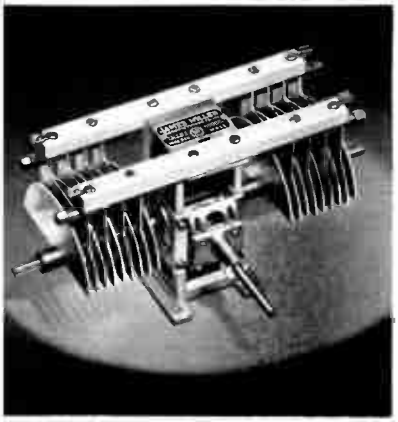
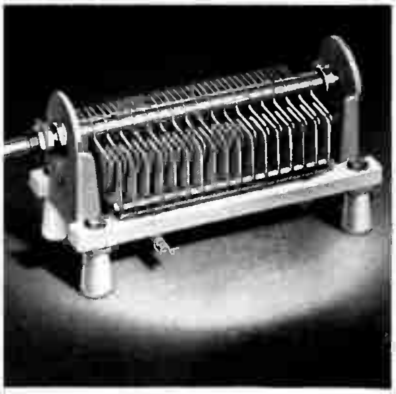


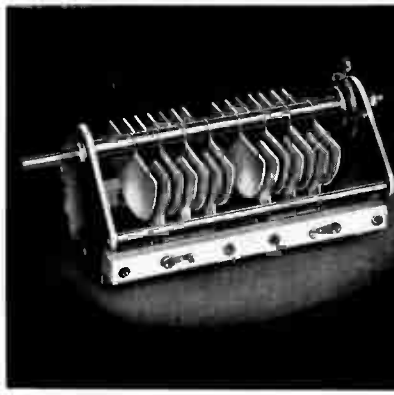
04000 and 11000 SERIES TRANSMITTING CONDENSERS

A new member of the "Designed for Application" series of transmitting variable air capacitors is the 04000 series with peak voltage ratings of 3000, 6000, and 9000 volts. Right angle drive, 1–1 ratio. Adjustable drive shaft angle for either vertical or sloping panels. Sturdy construction, thick, round-edged, polished aluminum plates with 1¾" radius. Constant impedance, heavy current, multiple finger rotor contactor of new design. Available in all normal capacities.

The 11000 series has 16 1 ratio center drive and fixed angle drive shaft.

Code	Volts	Capacity	Price
11035	3000	35	$ 6.90
11050	3000	50	7.14
11070	3000	70	7.80
04050	6000	50	16.00
04060	9000	60	18.00
04100	6000	90	18.00
04200	3000	205	20.00

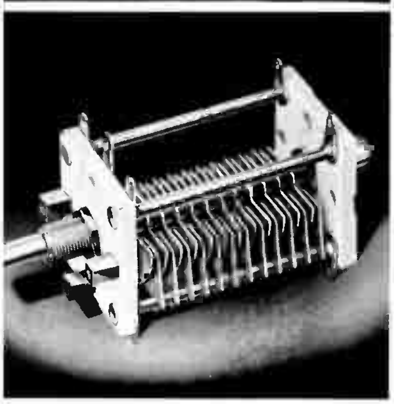


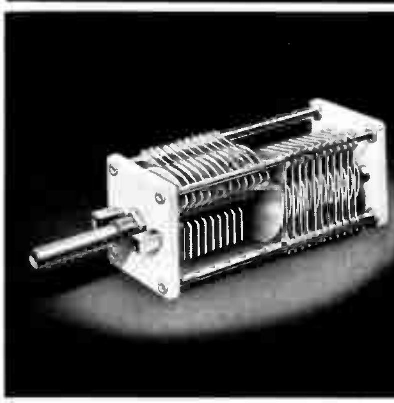


12000 and 16000 SERIES TRANSMITTING CONDENSERS

Rigid heavy channeled aluminum end plates Isolantite insulation, polished or plain edges. One piece rotor contact spring ond connection lug. Compact, easy to mount with connector lugs in convenient locations. Same plate sizes as 11000 series above.

The 16000 series has same plate sizes as 04000 series. Also has constont impedance, heavy current, multiple finger rotor contactor of new design. Both 12000 and 16000 series available in single and double sections and many capacities and plate spacing.





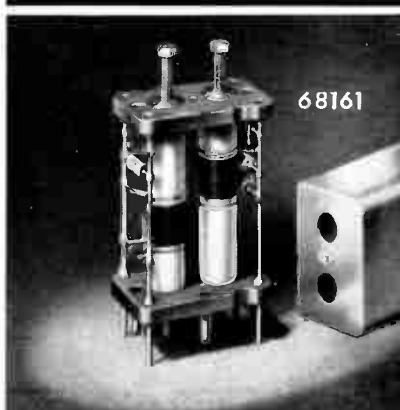
THE 28000–29000 SERIES VARIABLE AIR CAPACITORS

"Designed for Application," double bearings, steatite end plates, cadmium or silver plated brass plates. Single or double section .022" or .066" air gap. End plate size: 19/16" x 11/16". Rotor plate radius: ¾". Shaft lock, rear shaft extension, special mounting brackets, etc., to meet your requirements. The 28000 series has semi-circular rotor plate shape. The 29000 series has approximately straight frequency line rotor plate shape. Prices quoted on request. Many stock sizes.

NEUTRALIZING CAPACITOR

Designed originally for use in our own No. 90881 Power Amplifier, the No. 15011 disc neutralizing copacitor has such unique features as rigid channel frame, horizontal or vertical mounting, fine thread over-size lead screw with stop to prevent shorting and rotor lock. Heavy rounded-edged polished aluminum plates are 2" diameter. Glazed Steatite insulation.

No. 15011 .... $3.15



I.F. TRANSFORMERS

The Millen "Designed for Application" line of I.F. transformers includes air condenser tuned, and permeability tuned types for all applications. Standard stock units are for 456, 1600 and 5000 kc. B.F.O. also available.

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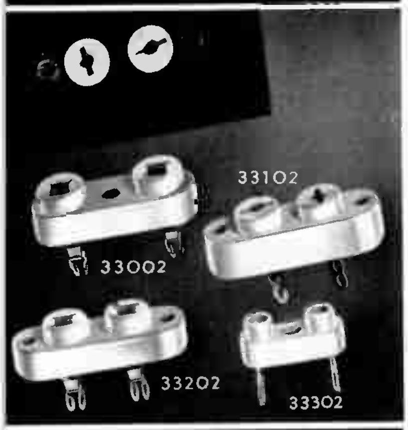
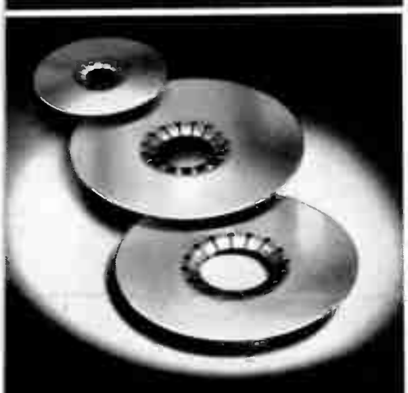
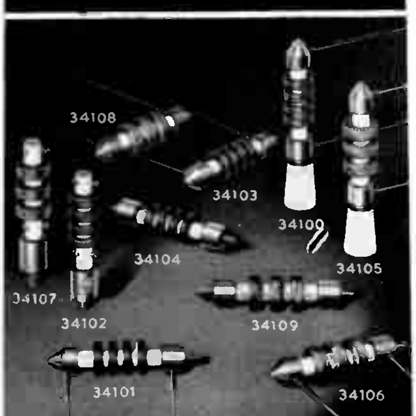






TUBE SOCKETS DESIGNED FOR APPLICATION

MODERN SOCKETS for MODERN TUBES! Long Flashover path to chassis permits use with transmitting tubes, 866 rectifiers, etc. Long leakage path between contacts. Contacts are type proven by hundreds of millions already in government, commercial and broadcast service, to be extremely dependable. Sockets may be mounted either with or without metal flange. Mounts in standard size chassis hole. All types have barrier between contacts and chassis. All but octal and crystal sockets also have barriers between individual contacts in addition.

The No. 33888 shield is for use with the 33008 octal socket. By its use, the electrostatic isolation of the grid and plate circuits of single-ended metal tubes can be increased to secure greater stability and gain.

The 33087 tube clamp is easy to use, easy to install, effective in function. Available in special sizes for all types of tubes. Single hole mounting. Spring steel, cadmium plated.

Cavity Socket Contact Discs, 33446 are for use with the "Lighthouse" ultra high frequency tube. This set consists of three different size unhardened beryllium copper multi-finger contact discs. Heat treating instructions forwarded with each kit for hardening after spinning or forming to frequency requirements.

Voltage regulator dual contact bayonet socket, 33991 black Bakelite insulation and 33992 with low loss high leakage mica filled Bakelite insulation.

No. 33004	$ .30
No. 33005	.30
No. 33006	.30
No. 33007	.34
No. 33008	.30
No. 33888	.18
No. 33087	.30
No. 33002	.30
No. 33102	.30
No. 33202	.30
No. 33302	.21
No. 33446*	5.00
No. 33991	.45
No. 33992	.55

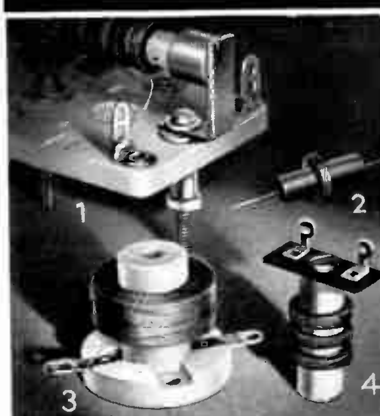
* For set of 3. Single discs $2.00 each.

RF CHOKES

Many have copied, few have equalled, and none have surpassed the genuine original design Millen Designed for Application series of midget RF Chokes. The more popular styles now in constant production are illustrated herewith. Special styles and variations to meet unusual requirements quickly furnished.

General Specifications: 2.5 mH, 250 mA for types 34100, 34101, 34102, 34103, 34104, and 1 mH, 300 mA for types 34105, 34106, 34107, 34108, 34109.

No. 34100	$.42
No. 34101	.36
No. 34102	.42
No. 34103	.36
No. 34104	.42



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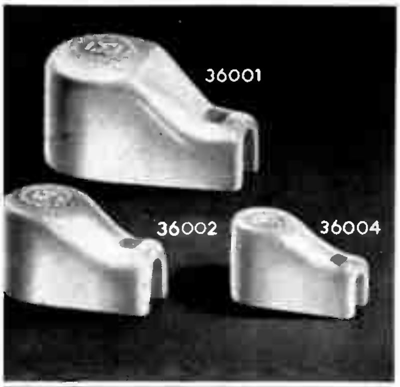
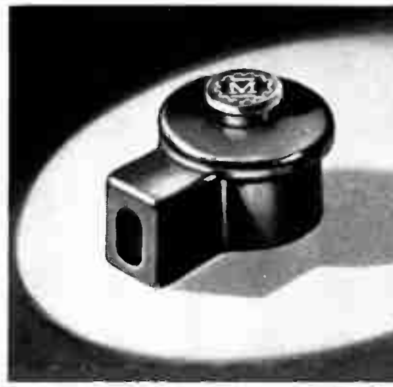




CERAMIC PLATE OR GRID CAPS

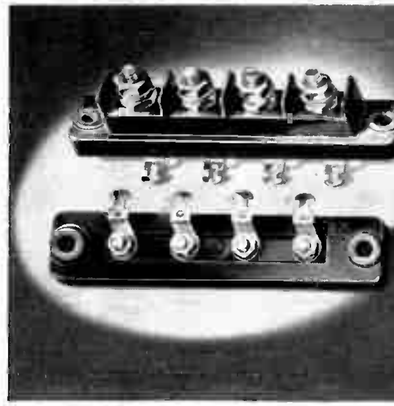
Soldering lug and contact one-piece. Lug ears annealed and solder dipped to facilitate easy combination "mechanical plus soldered" connection of cable.

No. 36001—9/16" . . . . $.21
No. 36002—3/8" 21
No. 36004—1/4" 21

SNAP LOCK PLATE CAP

For Mobile, Industrial and other applications where tighter than normal grip with multiple finger 360° low resistance contact is required. Contact self-locking when cap is pressed into position. Insulated snap button at top releases contact grip for easy removal without damage to tube.

No. 36011—9/16" . . . . $.60
No. 36012—3/8" 60



SAFETY TERMINAL

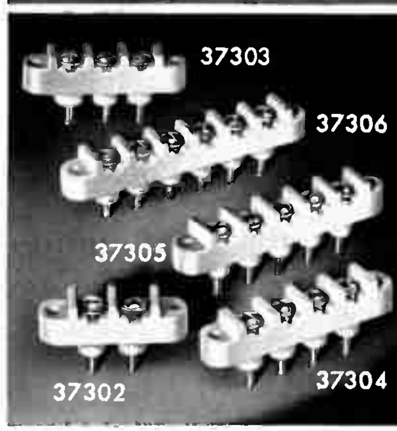
Combination high voltage terminal and thru-bushing. Tapered contact pin fits firmly into conical socket providing large area, low resistance connection. Pin is swivel mounted in cap to prevent twisting of lead wire.

No. 37001, Black or Red . . . . $.40
No. 37501, Low loss 55

TERMINAL STRIP

A sturdy four-terminal strip of molded black Textolite. Barriers between contacts. "Non turning" studs, threaded 8-32 each end.

No. 37104 . . . . $.60



POSTS, PLATES and PLUGS

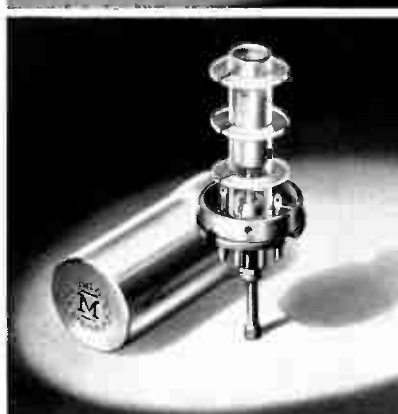
Designed for Application! Compact, easy to use. Made in black and red regular bakelite as well as low loss brown mica filled bakelite or steatite for R.F. uses. Posts have captive head.

No. 37202 Plates (pr.) . . . . $.30
No. 37212 Plugs 70
No. 37222 Posts (pr.) 40

STEATITE TERMINAL STRIPS

Terminal and lug are one piece. Lugs are Navy turret type and are free floating so as not to strain steatite during wide temperature variations. Easy to mount with series of round holes for integral chassis bushings.

No. 37302 . . . . $ .60
No. 37303 70
No. 37304 80
No. 37305 90
No. 37306 . . . . 1.00



MIDGET COIL FORMS

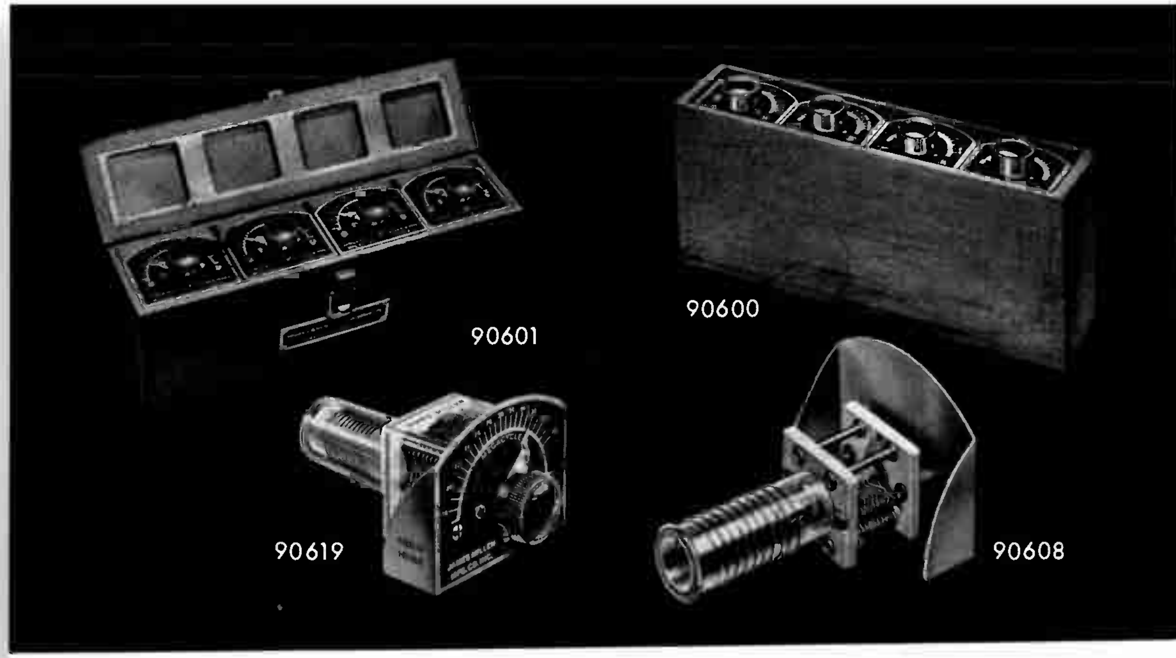
Made of low loss mica filled brown bakelite. Guide funnel makes for easy threading of leads through pins.

No. 45000 . . . . $.35
No. 45004 45
No. 45005 45

TUNABLE COIL FORM

Standard octal base of low loss mica-filled bakelite, polystyrene 1/2" diameter coil form, heavy aluminum shield, iron tuning slug of high frequency type, suitable for use up to 35 mc. Adjusting screw protrudes through center hole of standard octal socket.

No. 74001, with iron core $1.85
No. 74002, less iron core 1.50



Midget Absorption Frequency Meters

Many amateurs and experimenters do not realize that one of the most useful "tools" of the commercial transmitter designer is a series of very small absorption type frequency meters. These handy instruments can be poked into small shield compartments, coil cans, corners of chassis, etc., to check harmonics; parasitics; oscillator-doubler, etc., tank tuning; and a host of other such applications. Quickly enables the design engineer to find out what is really "going on" in a circuit.

Types 90605 thru 90609 are extremely small and designed primarily for engineering laboratory use where they will be handled with reasonable care. The most useful combination being the group of four under code No. 90600 and covering the total range of from 3.0 to 140 megacycles. When purchased in sets of four under code No. 90600 a convenient carrying and storage case is included. Series 90601 are slightly larger and very much more rugged. They are further protected by a contour fitting transparent polystyrene case to protect against damage and dirt. This latter series is designed primarily for field use and are not quite as convenient for laboratory use as the 90605 thru 90608 types. All types have dials directly calibrated in frequency.

Code	Description	Net Price
90604	Range 160 to 210 mc.	$ 6.00
90605	Range 3.0 to 10 mc.	4.50
90606	Range 9.0 to 23 mc.	4.50
90607	Range 23 to 60 mc.	4.50
90608	Range 50 to 140 mc.	4.50
90609	Range 130 to 170 mc.	6.00
90610	Range 105 to 150 mc.	6.00
90619	Range 350 to 1000 kc.—Neon Indicator	15.00
90620	Range 150 to 350 kc.—Neon Indicator	15.00
90625	Range 2 to 6 mc.—Neon Indicator	15.00
90626	Range 5.5 to 15 mc.—Neon Indicator	15.00
90600	Complete set of 90605 thru 90608, in case	18.00
90601	Complete set Field type Frequency Meters in metal carrying case 1.5 to 40 mc.	29.00

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150 EXCHANGE ST., MALDEN, MASSACHUSETTS, U.S.A.

JK CRYSTALS

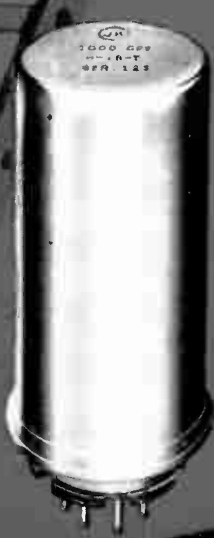
Crystals

FOR THE CRITICAL

There are no rungs missing in the JK kilocycle ladder!

No matter how special your crystal needs may be, the JAMES KNIGHTS CO. is fully equipped to meet them. There's a complete selection for commercial, broadcast, industrial and amateur applications. Special crystals to fit special needs will be built to order at modest price. Your inquiries are invited.

Type H17
120 MC



Type H18T
3000 Cycles

Two James Knights Co. planes are available for the handling of emergencies

THE JAMES KNIGHTS CO.
Sandwich, Illinois
Stabilized Quartz Crystals

1950 Parade of Collins Stars

COLLINS 51J-1 COMMUNICATIONS RECEIVER

The Collins 51J-1 is a double conversion superheterodyne, permeability tuned throughout, and continuously tunable over the range 0.5-30.5 megacycles. Designed as a general purpose communications receiver for military, commercial and individual use, the 51J's outstanding characteristics are extremes of accuracy and stability. Quartz crystals in the first conversion circuit, and the very accurate, stable Collins 70E-7A VFO in the second conversion circuit, contribute notably to these characteristics.

The tuning method employed is an innovation. The range is divided into 30 bands of 1,000 kc each. The tuning mechanism is based on a decade system in which the megacycle figure is set by means of a band switch. 100 kc figures are indicated on the slide rule dial; kilocycle figures on the circular dial. Under normal operating conditions, with a 10-minute warmup, the dial reading is within 2 kc of the receiver's exact frequency throughout its range. Calibration error can be reduced to less than 200 cycles by means of an adjusting knob which permits the reading to be corrected at 100 kc intervals by use of a built-in crystal oscillator. The 100 kc crystal may be calibrated directly against WWV.

Even without reference to the crystal calibrator, the frequency over the temperature range −20° C to +60° C does not vary from the frequency at 20° C by more than 30 parts per million plus 1 kc; thus stability is within 2 kc at the highest operating frequency. Frequency does not vary more than 100 cycles from that at 115 line volts when this voltage is varied through the range 105 to 125. Changes in atmospheric pressure from sea level to 10,000 feet altitude, relative humidity from 10 to 90%, and mild shock, do not vary the frequency by more than 500 cps.

The 51J is constructed in a panel and shelf assembly for standard rack cabinet mounting, and is protected by a dust cover. It can be supplied optionally in a table mounting cabinet.

Dimensions: Panel 19" wide, 10½" high, 13" behind the panel.
Cabinet 21⅛" wide, 12⁷⁄₁₆" high, 13⅞" deep.

Power source: 115 volts 50/60 cycles a-c.

Weights: Receiver 35 pounds, cabinet 20 pounds.

Net domestic price, rack panel or cabinet mounting, complete with tubes, speaker and matching cabinet assembly, and instruction book (including excise tax but exclusive of any state tax)..$875.00

For results in amateur radio, it's . . . COLLINS

COLLINS RADIO COMPANY, Cedar Rapids, Iowa

11 W. 42nd St.	2700 West Olive Ave.	M & W Tower	Fountain City Bank Bldg.
NEW YORK 18	BURBANK	DALLAS 1	KNOXVILLE

Collins 32V-2 Transmitter

Input: 150 watts c-w, 120 watts phone

The 32V-2 is a VFO controlled bandswitching, gang-tuned amateur transmitter, conservatively rated at 150 watts input on c-w and 120 watts input on phone. It covers the 80, 40, 20, 15, 11 and 10 meter bands. The entire transmitter, including power supply, is cased in a cabinet identical in size to that of the 75A-1 receiver.

The heart of the 32V-2 is the Collins 70E-8A permeability tuned oscillator, used as the VFO. The frequency range of the 70E-8A, 1600-2000 kc, is covered in 16 turns of the vernier dial. The calibration is very accurate, and stability compares favorably with most crystals used by amateurs. To assure freedom from humidity effects, this oscillator is baked dry, then completely sealed and moisture proofed. As an added protection, a silica gel capsule is inserted in the oscillator at the factory.

The r-f tube line-up: a 6SJ7 VFO, 6AK6 buffer, 6AG7, 7C5 and 7C5 frequency multipliers, and 4D32 final amplifier. Speech line-up: a 6SL7 in cascade to a 6SN7 to a pair of 807 modulators, which furnish 60 watts audio power to modulate the final amplifier. The power supply contains a 5Z4 (low voltage) and two 5R4GY (high voltage) rectifiers, a VR-75 bias regulator, and two 0A2 screen voltage limiters.

All controls are conveniently located on the front panel. As an additional refinement, both coarse and fine antenna loading controls are actuated by the same dial. The 32V-2 can be operated by a push-to-talk switch on the microphone, a key, or a separate switch.

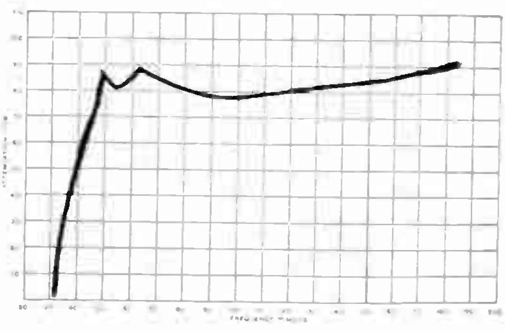
Terminals are provided for supplying the energizing voltage for an antenna change-over relay. Other terminals, paralleled with the operate switch, are used to disable the receiver when the transmitter is in SEND position. Grid-block keying is utilized on three stages following the VFO. The back-wave of the VFO as heard in a receiver placed beside the 32V-2 is negligible; thus break-in operation is accomplished without difficulty. Keying is very clean, without chirp or clicks. The keyer circuit also includes a side-tone oscillator which is used as a c-w keying monitor.

Dimensions: 21⅛" wide, 12⁹⁄₁₆" high, 13⅞" deep.
Power source: 115 volts 50/60 cycles a-c.
Shipping weight: 128 pounds.
Net domestic price, complete with tubes and instruction book, $575.00.

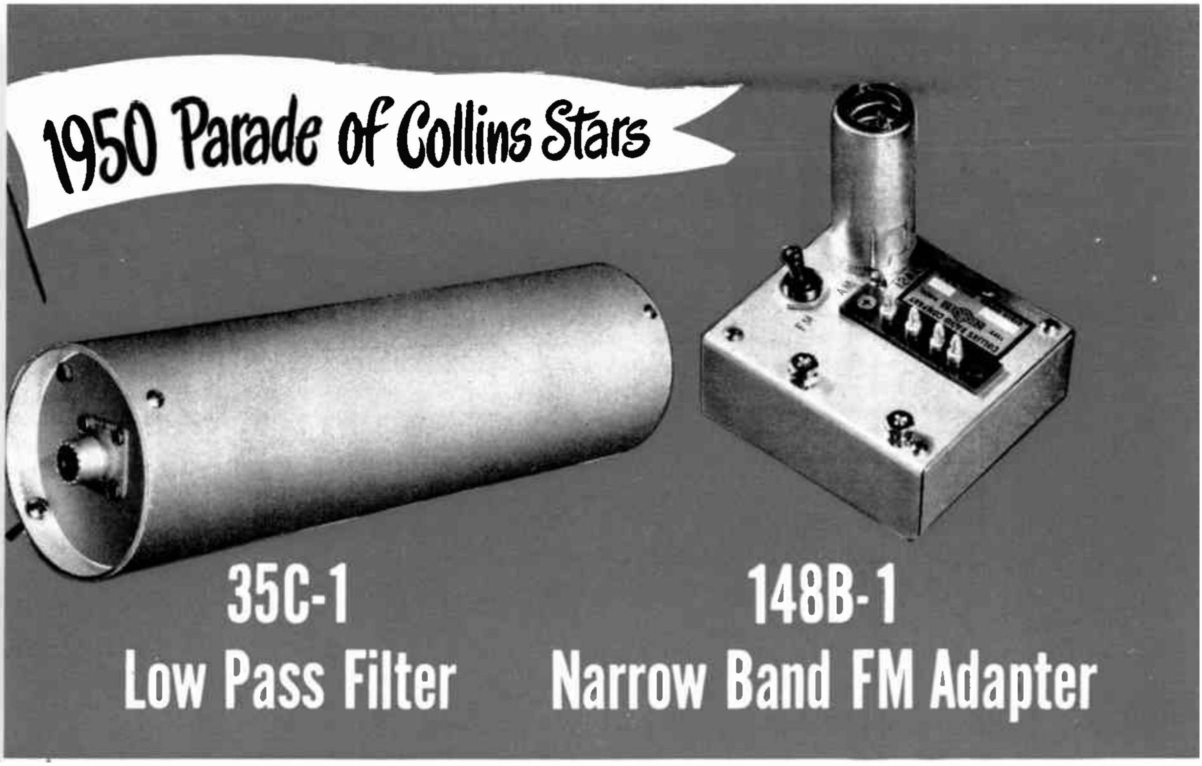
TVI Reduction—The following methods of avoiding TVI have been provided in the design of the 32V-2 and accessory units:

(a) Reduction of spurious signals in the transmitter output.
(b) Filtering of transmitter output at the antenna terminal.
(c) Shielding of transmitter.

(a) In the 32V-2 series added tuned circuits in the exciter and an added L section in the unbalanced pi output network reduce unwanted signals. This output network is designed primarily to feed into a 52 ohm coaxial transmission line, such as RG-8/U. It will also match unbalanced impedances of approximately 26 to 300 ohms and will tune out reactances normally encountered.



35C-1 *low pass filter attenuation curve*

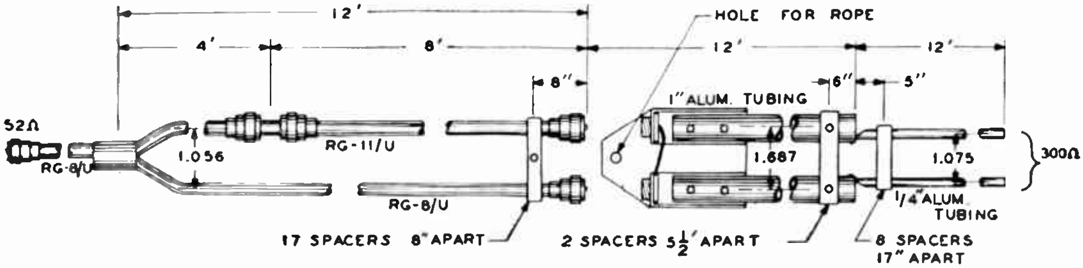


35C-1 Low Pass Filter (b) A coaxial fitting is provided at the rear of the 32V-2 cabinet. This permits the use of a well shielded transmission line in which the Collins 35C-1 Low Pass Filter may be inserted. The 35C-1 is a 52 ohm three-section filter which, with approximately 0.2 db insertion loss below 29.7 mc, provides approximately 75 db attenuation of harmonic emissions at the television frequencies (see attenuation curve). This high attenuation is added to that provided in the transmitter. The unbalanced output permits grounding of the outer conductor of the line and the case of the filter. The price of the 35C-1 is $40.00 at your Collins dealer's.

49S-1 Shielded Cabinet (c) For reducing TVI from sources other than the antenna, the Collins 49S-1 Shielded Cabinet for the 32V-2 is available at extra cost. It includes well filtered control wires and leads to terminals, and forced air ventilation. Provision is made for mounting the 35C-1 filter on the rear. The 49S-1 Cabinet is required in only the most difficult TVI installations. If wanted, your new 32V-2 can be delivered in the 49S-1 by your dealer. Or, if you already own a 32V-2, you can order from him a 49S-1 cabinet only, and install it yourself.

315E-1 Balun Transformer—For best operation, the 35C-1 filter should feed a properly terminated 52 ohm line. Coupling to a balanced antenna may be accomplished by an antenna tuner or by the Collins 315E-1 Balun Transformer, which is a wide band, low loss transmission line (see diagram) for coupling from a 52 ohm unbalanced line to a 300 ohm balanced load without tuning controls. It consists of a transmission line connected to transfer from unbalanced to balanced conditions ("balun") and a step-tapered impedence matching line. Over the frequency range 7 to 30 mc, a standing wave ratio of less than 2 to 1 is possible. The efficiency of the system is good even beyond the specified limits. The 315E-1 is supplied in kit form with coaxial cables completely made up, and aluminum tubing and spacers fabricated ready to assemble. Available through your Collins dealer.

148B-1 Narrow Band FM Adapter—The Collins 148B-1 Narrow Band FM Adapter is for use with either the 32V-1 or the 32V-2 amateur transmitters. It plugs into the 70E-8 variable frequency oscillator, and is suitable for FM operation on all bands. Frequency deviation is adjusted by the Audio Gain control on the transmitter. A toggle switch selects AM or FM.



315E balun transformer schematic

COLLINS 75A-1 AMATEUR RECEIVER

The well known and highly regarded 75A-1 receiver was designed specifically to give the radio amateur the best possible performance in the 80, 40, 20, 15, 11 and 10 meter bands.

Double conversion and crystal filter controls, with a high frequency first i-f and a low frequency second i-f, provide at least 50 db image rejection in all bands. The received bandwidth is variable in 5 steps from 4 kc to 200 cycles at 6 db down from the peak of the resonant frequency. The 6AK5 r-f stage makes possible a threshold sensitivity far better than can be realized in normal installations.

Very high accuracy and stability result from the use of precision quartz crystals in the first conversion circuit, the extreme accuracy and stability of the v.f.o. in the second conversion circuit, and linearity and absence of backlash in the tuning mechanism. The bandlighted slide rule dial indicates frequency in megacylces, while the vernier dial provides a direct reading in kilocycles. Panel controls include tuning, bandswitch, r-f gain, audio gain, c-w pitch, on-off-standby, crystal selectivity, crystal phasing, avc-manual-c-w, and noise limiter switch.

Dimensions: 21⅛" wide, 12 7/16" high, 13⅞" deep.
Power source: 115 volts 50/60 cycles a-c.
Shipping weight: 93 pounds including speaker.
Net domestic price, complete with 13 tubes and rectifier, speaker and cabinet assembly, and instruction book (exclusive of state tax but including excise tax)........................$375.00

The New One Kilowatt Transmitter . . . Available Early Summer, 1950

Engineered by Collins expressly for radio amateurs. One kilowatt input, both AM phone and c-w. Designed to avoid TVI. Bandswitching throughout. Usual excellent Collins audio.

Tentative specifications: Exciter, single control tuning. Dial similar to the new Collins 51J-1 receiver. Output 50 to 75 ohms; single ended pi followed by L section. Tubes: P. A., two type 4-250A; modulator, two type 810.

FOR HAMS, ENGINEERS

IT'S SYLVANIA TUBES

CATHODE RAY TUBES

More than 75% of leading television set manufacturers use Sylvania Television Picture tubes. You'll want Sylvania cathode ray tubes, too, for television replacement, industrial or experimental use. Both magnetic and electrostatic types available in screen sizes from 2 to 16 inches. Tubes have excellent brightness and definition and external conductive coating, which, when grounded, acts as a filter capacitor.

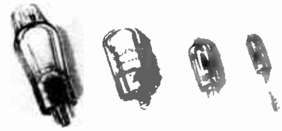
OSCILLOSCOPES

Two unique push-pull amplifiers give extra sensitive patterns on the 7GP1 screen of the Sylvania Type 132 Oscilloscope (shown). One inch peak-to-peak vertical deflection is obtained with .21 rms input. Z-axis input provided for intensity modulation applications.

Type 131 (3-inch screen) utilizes a type 3AP1 cathode ray tube having electrostatic deflection and focus. With horizontal and vertical amplifiers 0.5 volt rms gives 1-inch peak-to-peak deflection. Sweep generator variable from 15 to 40,000 cycles.

MODULATION METER

Now you can monitor your modulation percentage and speech quality with a fine instrument at low cost—Sylvania Modulation Meter Type X-7018. Percentage modulation can be read directly on the meter. Headphone jack also permits monitoring your signal quality to check for hum or audible distortion. Indicates carrier shift.



RECEIVING TUBES

You'll want Sylvania tubes—from the tiny subminiatures to the famous Lock-In—for their well-known quality and performance. Sylvania has its own plant for the manufacture of radio tube parts. Every step is quality-controlled. You can be sure of satisfaction when you buy any tubes from the big Sylvania Receiving Tube line.

TRANSMITTING TUBES

Sylvania's comprehensive line of quality transmitting tubes consists of more than 20 different types . . . triodes, beam power tubes and mercury vapor and vacuum diodes.

Typical of this complete line —designed and built to the exacting standards that have made Sylvania Receiving Tubes the leaders—is the 2E26 (shown) . . . a pentode rf amplifier and oscillator. Sylvania Transmitting Tubes are also available for service as af power amplifiers and modulators, and as rectifiers.

GERMANIUM DIODES

In addition to Sylvania's big line of ceramic-type germanium crystal diodes, duodiodes and varistors for a multitude of AM, FM and television applications... Sylvania Electric now offers smaller, lighter germanium diodes hermetically sealed in glass! This construction makes diodes moisture-proof, gives greater electrical stability . . . they're ideal for side-by-side mounting, no risk of accidental contact. Types 1N34A and 1N58A are immediately available.

SYLVANIA

RADIO TUBES; CATHODE RAY TUBES; ELECTRONIC DEVICES; FLUORESCENT LAMPS

AND TEST EQUIPMENT!

AUDIO OSCILLATOR

One of the most versatile and convenient test instruments made, the Sylvania Audio Oscillator Type 145 provides a powerful, accurate tone source for distortion checking of radio receivers.

It may be used as a modulating signal for radio transmitters or as a simple frequency meter. It is ideal for response and distortion testing of audio amplifiers, public address systems, juke boxes, wired musical installations and individual speaker cones. Frequency range 20 to 20,000 cycles. Output impedance variable by 3-position panel selector switch.

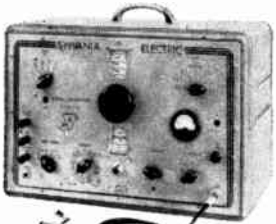
POLYMETER

The Sylvania Polymeter Type 221 is a multi-purpose vacuum tube voltmeter that greatly simplifies the job of applying many accurate measurements and tests to radio and television equipment. Electrical values measured include audio, ac and rf voltages (up to 300 mc); dc voltages from 0.1 to 1,000; direct currents from .05 milliampere to 10 amperes; resistances from ½ ohm to 1,000 megohms.

New plus features for complete television service: 1. shielded ac probe lead—reduces stray field effects; 2. microphone type panel connectors on probe leads insure firm long life connections; 3. RF probe features ground clip and detachable extension tip—extremely flexible in application.

VOLTAGE MULTIPLIER PROBES

With these two DC Voltage Multipliers, the 1,000 vdc range on your Sylvania Polymeter will extend to 10,000 or 30,000 vdc full scale. Add either of these accessories to your Polymeter and you have a Kilovoltmeter for testing TV circuits and other high dc voltage applications. Types 222 and 223 are 10 KV Probes; Types 224 and 225, 30 KV Probes.



SIGNAL GENERATOR

With Sylvania's new Signal Generator Type 216 you can align the rf and if sections of all FM and AM receivers, adjust all types of FM detectors, and make overall receiver checks. Its high level output and accurate calibrations make it also a valuable instrument for other service and laboratory uses requiring a high quality rf signal source.

MAIL THIS COUPON TODAY!

SYLVANIA ELECTRIC PRODUCTS INC.
Radio Tube Division, Advertising Dept.
Emporium, Pa.

Gentlemen: Kindly forward information on items checked.

☐ Receiving Tubes ☐ Cathode Ray Tubes ☐ Transmitting Tubes ☐ Oscilloscopes ☐ Signal Generator ☐ Polymeter ☐ Voltage Multiplier Probes ☐ Audio Oscillator ☐ Germanium Diodes ☐ Modulation Meter

NAME ______

ADDRESS ______

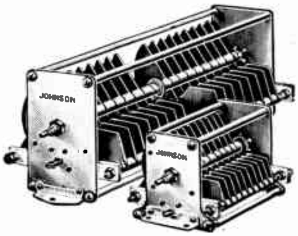
CITY ______

STATE ______ ZONE ______

ELECTRIC

-IXTURES, WIRING DEVICES, SIGN TUBING; LIGHT BULBS; PHOTOLAMPS

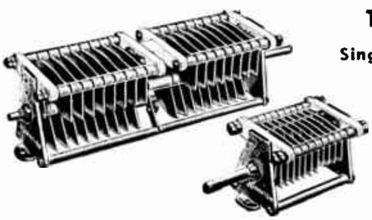
JOHNSON VARIABLE CONDENSERS



TYPES C and D

Single and Dual Models

Sturdily constructed to give trouble-free operation under the most severe service. High in quality yet low in price. Heaviest aluminum plates of any similar condenser, .051″ thick. Steatite insulation—large laminated rotor brushes—heavy 5/16″ diameter aluminum tie rods for frame strength and rigidity—¼″ cadmium-plated steel shafts. Mounting brackets fit either top or bottom of end plate so that stators may be mounted either top or bottom. Panel space, Type C, 5½″ wide x 5⅜″ high—Type D, 4¼″ wide x 4″ high. Available in capacities, from 12 to 496 mmfd., and voltage ratings from 3500 V. to 11,000 V.



TYPES E and F

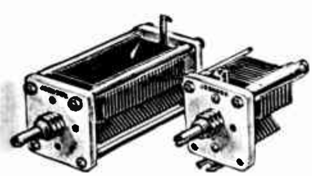
Single and Dual Models

Designed as rugged, compact units for medium and low power transmitters, type E and F condensers are in a class by themselves. They have more capacity per cubic inch and occupy less panel space for their rating than any other condenser.

Heavy aluminum plates, .032″ thick, with rounded edges for maximum voltage rating—Heavy aluminum tie rods ¼″ diameter for frame strength and rigidity—Steatite insulation—heavy phosphor bronze contact springs, cadmium plated—Center contact on dual models—Chassis or panel mounting—Stainless steel shafts.

In addition to mounting foot shown, removable single hole brackets are furnished so that condenser may be inverted from position shown, or other components mounted above.

Panel space, Type E, 2⅝″ wide x 2 19/32″ high. Panel space, Type F, 2 1/16″ wide x 2″ high. Available in capacities, from 7 to 488 mmfd., and voltage ratings from 2000 V. to 4500 V.



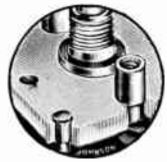
TYPE H CONDENSER

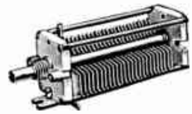
Single Section with single or double end plates and Dual Section Models

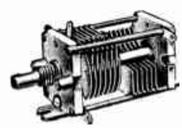
The type H condenser was designed for aircraft transmitters and combines a minimum of weight and size with simple but rugged construction. Use of steatite for end plates permits panel mounting with both rotor and stator insulated from ground. Has aluminum plates .020″ thick. End plate 1 ½″ square. Available in a wide range of capacities. Voltage ratings of 1500 V. and 3000 V.

TYPE G CONDENSER

The Type G condenser is extremely popular as a neutralizing condenser for medium and low power stages. It is also widely used for grid and plate tuning at high and ultra-high frequencies. It has a single end plate of steatite and low minimum capacity. .032″ rounded aluminum plates, and front and rear shaft extension are among outstanding features. Available in capacities from 3.5 to 52 mmfd., and voltage ratings of 2000 V. to 7000 V.







TYPE L CONDENSER

Single — Dual — Butterfly-Differential
Ceramic Soldered

Outstanding feature is the use of perfected ceramic soldering which assures absolute—and permanent—rigidity and strength, absolute—and permanent—maintenance of capacities!

There are no eyelets, nuts or screws to work loose causing stator wobble and fluctuations in capacity. JOHNSON ceramic soldering leaves a bond which is stronger than the rugged steatite end plates themselves.

Silent operation on the highest frequencies is assured with a split sleeve tension bearing that also prevents fluctuations in capacity.

Ideal for peak efficiency even under the severest conditions, such as portable-mobile operation.

Available in capacities from 2.8 to 202 mmfds., and voltage ratings of 1500 V. and 3000 V.

TYPE M MINIATURE

The smallest air variables ever built. A necessity in high frequency equipment. Available in single, differential and butterfly types. Single hole mounting. Split sleeve rotor bearings—no shaft wobble. Steatite end frames. Panel mounting space is ¾″ by ⅝″ Capacities from 1.5 mmfd. to 19.7 mmfd., and voltage rating of 1250 V.

TYPE J CONDENSER

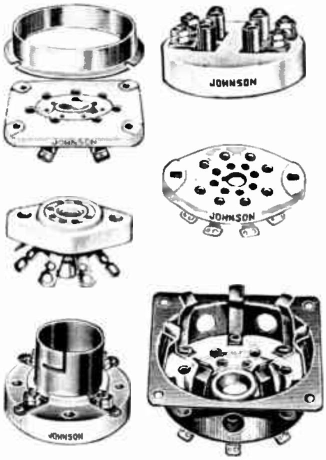
The Type J condenser is a midget with big condenser characteristics. It has wider spacing than most small types, yet occupies little more space and is ideal for oscillator and low power stages. The spacing is .025″ and universal type mounting brackets make possible a variety of mountings including chassis, panel, or inside tube socket type inductors. Steatite end plate is 1⅛″ wide. Capacities from 2.6 mmfd. to 102 mmfd., and voltage rating of 1200 V.

TYPE N NEUTRALIZING

Small mounting space requirements, extremely high voltage rating in proportion to size, fine adjustment with uniform voltage breakdown rating throughout the full capacity range, and low cost, make these neutralizing condensers ideal for the modern transmitter. "Plates" are aluminum cups supported on a steatite frame with cast aluminum mounting bracket. Because of the design these condensers will withstand much higher voltage than conventional flat plate condensers of the same spacing. Capacities from 1.1 mmfd. to 11.0 mmfd. and voltage ratings of 8500 V. to 14,500 V.

E. F. JOHNSON COMPANY, WASECA, MINN., U. S. A.

JOHNSON RADIO TRANSMITTING COMPONENTS



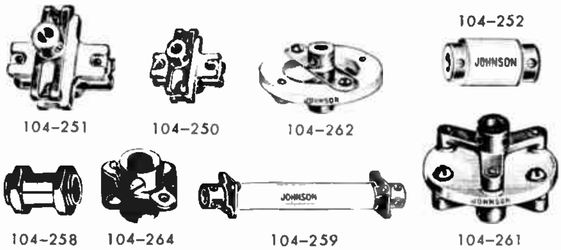
TUBE SOCKETS

Johnson tube sockets are manufactured in a wide range of styles and sizes to meet every need. Available are industrial bayonet sockets, miniature sockets, acorn sockets, wafer sockets and special types.

The industrial bayonet sockets have heavy metal shell for strength and adequate insulation for high voltage applications. Johnson wafer sockets are all insulated with grade L-4 steatite or better, top and sides are glazed. Johnson acorn sockets have contacts of silver plated beryllium copper with base grade L-4 steatite.

The Johnson "Tube Socket Guide" is available on request.

SHAFT COUPLINGS



All Johnson insulated shaft couplings are characterized by steatite insulation properly proportioned for electrical and mechanical strength, and by accurate metal parts heavily plated.

The phosphor bronze springs of the —250 and —251 series couplings provide flexibility without backlash and compensate for minor shaft misalignments. Rigid types —252, —262 and —261 meet the requirements of accurate shaft alignment and high torque.

The —259 and —2593 are bar type couplings recommended for high voltages or very high frequencies.

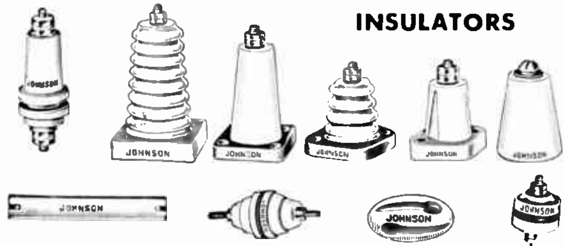
The —264 is a small bakelite insulated flexible coupling for DC or low voltage RF applications.

PILOT, DIAL and INDICATOR LIGHTS

Johnson dial and indicator light assemblies are outstanding examples of sound engineering design, excellent material and careful workmanship. Their use is your assurance of complete satisfaction.

Johnson carries a complete line of hundreds of standard pilot light assemblies to meet every ordinary need. Special assemblies, to meet specific requirements can be furnished in production quantities on special order. Your inquiries are invited.

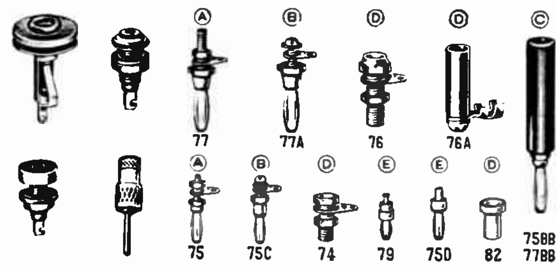
INSULATORS



Johnson insulators are specifically designed for high frequencies. Insulating materials were selected after exhaustive laboratory tests. Superior grade, low absorption, well glazed electrical porcelain, and Grade L-4 or better steatite are used.

In addition to fine quality insulating materials the Johnson line distinguishes itself with perfection of ceramic design; logical proportions; clean-cut, accurate molding; and high grade nickel-plated brass hardware, with milled (not stamped) nuts.

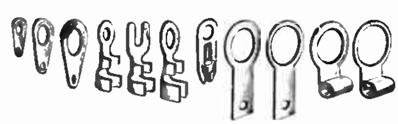
PLUGS and JACKS



Those who appreciate quality, right down to the last detail, will appreciate the superiority of Johnson plugs and jacks.

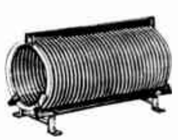
There is a wide assortment to choose from including banana spring types, spring sleeve types, plastic head tip jacks, molded round head tip jacks, insulated combination jacks, metal head tip jacks, twin tip jacks and shorting type twin tip jacks.

TINNED COPPER SOLDERING TERMINALS



Available in eleven sizes, Johnson soldering terminals meet the requirements of most applications. Composed of copper for low resistance they are tinned to permit easy soldering.

EDGEWISE WOUND "Hi-Q" INDUCTORS



Design improvements and mycalex insulation are features in this inductor of plated edge-wound copper strip. They are widely used in commercial equipment, and will safely handle more than 1000 watts in continuous service. Other sizes and types of inductors are manufactured for commercial broadcast and industrial electronic applications. More detailed information available on request.

WRITE FOR FREE GENERAL PRODUCTS CATALOG

JOHNSON *. . . a famous name in Radio*

E. F. JOHNSON CO. WASECA, MINNESOTA

TRANSMITTER TRIUMPH!

The JOHNSON

VIKING I KIT

150 Watts Input AM Phone and CW
Bandswitching 10-160 Meters

The Viking I offers hams commercial transmitter design, efficiency and appearance at kit prices. Superlative performance and operating convenience set new standards for amateur transmitters. A full 100 watts of AM phone or CW on 160, 80, 40, 20, 15 or 10 meters at your finger-tips.

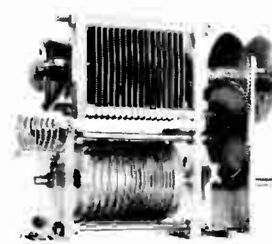
The pi-section output stage will efficiently load many antennas without external couplers. The final tank coil is a variable inductor with excellent insulation and high Q throughout its range. Plug-in coils are completely eliminated!

Novice or old timer can obtain brilliant performance from the Viking I. A wiring harness, punched chassis and panel, table cabinet, carefully detailed instructions and all parts furnished with the exception of tubes, crystals, mike and key.

AMATEUR NET $209.50

Features

- Amplitude Modulation
- Band Switching
- 100 Watts Phone Output
- 115 Watts CW Output
- 10 Crystal Positions
- Front Panel Controls
- VFO Input Receptacle
- VFO Power Outlet
- Two Complete Power Supplies
- Pi-Network Coupling
- All Stages Metered
- Self Contained



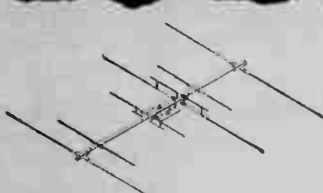
Final Tank Permits Continuous Tuning

TUBE LINE-UP

6AU6 crystal oscillator	6AU6 voltage amplifier	5R4 HV rectifiers
6AQ5 buffer doubler	6AU6 driver	5Z4 LV rectifier
4D32 final amplifier	807 pp modulators	6AL5 bias rectifier

The Ultimate in Beams.. the new universal JOHNSON ROTOMATIC

available with parasitic or driven elements



Designed for those who want the finest, here's the most flexible beam ever offered the ham. Parasitic arrays for 10, 15 or 20 meters, as well as dual beams for two of these bands. In addition, there are the new Johnson designed unidirectional phased arrays employing driven elements. These arrays, having the same gain and front to back ratio, can be erected and tuned without the usual laborious adjustment required by past beams.

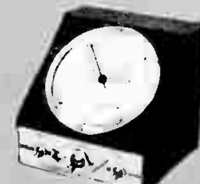
The elements, rotator, direction indicator, etc. may be purchased separately.

Write For New Rotomatic Folder

Rotator scoffs at sleet, high winds, turns on coldest mornings. Rotation reversible, 360° at 1¼ RPM.

Features Galore

- High Gain
- Excellent Front to Back Ratio
- Light Weight
- Efficient T Match
- New "Tippable" Rotator
- Will Safely Handle 2½ KW
- All-weather Construction
- Easy to Erect
- Availabe With Parasitic Or Unidirectional Phased Elements
- Dual Beams Available



Selsyn indicator, containing motor control and antenna relay.

Heavier Windings on New

JOHNSON HAM INDUCTORS

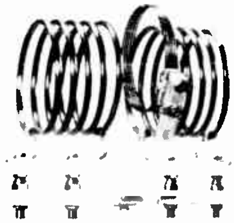
Match Tube to Coil — Link to Line
For Higher Efficiency, Harmonic Reduction

The Johnson "Air Wound Inductor Catalog" is loaded with facts — write for your free copy.

Better efficiency and harmonic reduction are secured through the availability of two fundamental types of inductors for each band, one type for high voltage low current tubes, the other for use with low voltage, high current tubes. They are available for all ham bands in 150, 500 and 1000 watt ratings employing HEAVIER windings.

In addition, there's a complete line of plug-in links, which Johnson pioneered for the above inductors.

Coils, jack bar assemblies, swinging link arms, etc. fit present day, competitive components — can be purchased separately.



Coil windings are a wire size LARGER.

Johnson was first with the plug-in link.

JOHNSON Q ANTENNAS

Matching Ease and Radiation Efficiency Never Surpassed!

There's a good reason for the ever rising popularity of the Johnson Q Antenna — matching ease and radiation efficiency have never been surpassed.

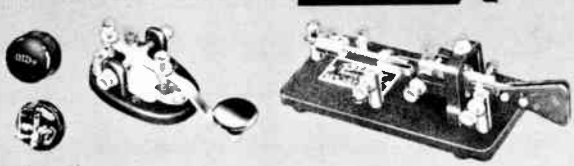
They're available for 2, 6, 10, 15, 20 and 40 meters. The 2Q and 6Q use aluminum tubing for the radiation portion, also. Applications for the Q are practically endless. The Johnson "Q" beam consists of two "Q's" spaced 1 5 wave employing two "Q's" for the lower frequency of the two bands desired.

New Johnson Antenna Handbook

Just off press. 47 pages of valuable information on antennas, transmission lines, coupling, etc. Sixth Edition

Price 60c

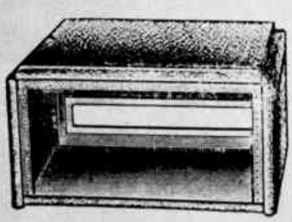
JOHNSON SPEEDX TRANSMITTING KEYS



Every cw man — beginner or expert — will quickly appreciate the smooth, snappy action of Johnson SpeedX keys. It encourages faster, better sending with less conscious effort. Many models to choose from, in either semi-automatic or heavy duty hand models.

RADIO CABINETS, CHASSIS, PANELS

Aluminum for Lightness • Steel for Strength



Johnson floor model cabinets feature unique adjustable rails for standard relay panels, recessed toe spaces at front and sides, inside ventilation and reversible rear door. Available in gray or black from 48-3/4" overall height to 68" panel space from 42" to 61-1/4". Size is 22" wide by 17-1/2" deep. Panel width 19".

Table model cabinets are of all aluminum construction .064" thick. Parts for attaching panels are double thickness, tapped for 10 — 32 screws. Overall height from 11-1/4" to 14-3/4", panel space from 8-3/4" to 12-1/4". Size 21" wide by 15" deep.

Table model cabinets with both top and rear door are equipped with positive flush snap-catch and may be installed to hinge from either side. Overall height 28-3/4", panel space 26-1/4". Size is 21" wide by 15" deep.

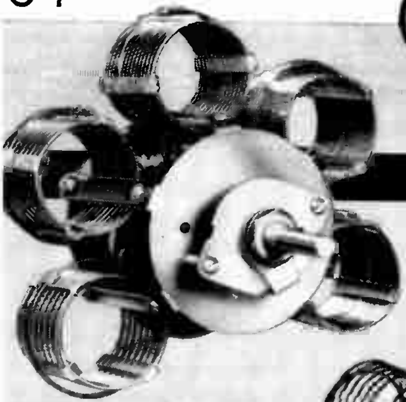
Relay rack panels are 1/8" thick aluminum for lightness and easy working. Panels are 19" long. Chassis are die cut and have no overlaps at corners.

JOHNSON . . . *a famous name in Radio!*

E. F. JOHNSON CO., WASECA, MINNESOTA

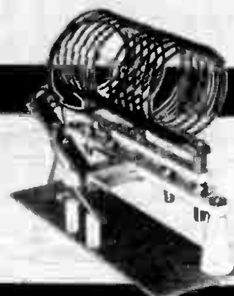
SEE NEXT PAGE

B&W PARTS and



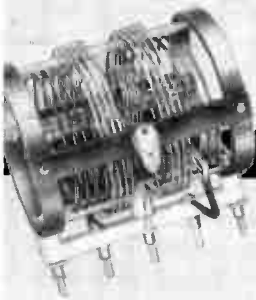
B & W BAND SWITCHING TURRETS

Pioneered by B & W—These compact 5-band switching Turrets are available from 35-watt ratings up through units that may be operated at voltages up to 1000 volts and input powers as high as 150 watts. They provide instant band switching, covering the 80 to 10 meter bands, and are regularly stocked as center-linked, center-tapped coils or end-linked, untapped.



B & W TYPE HD INDUCTORS

Rated up to 1000 Watts Input—Three general types available: without link, fixed center link with center tap, and variable center link with center tap. Type HD coils are ruggedly built, reasonably priced and are typical of the many B & W coils available for specific applications.

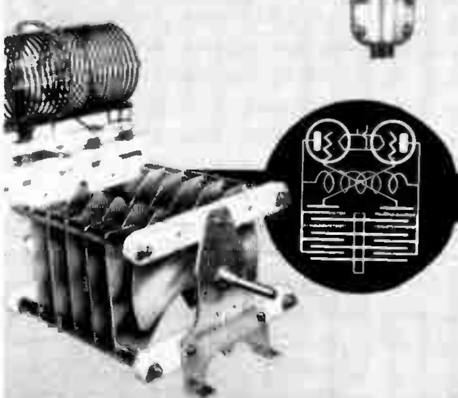


B & W 3400 SERIES INDUCTORS

With Fixed and Adjustable Coupling—Designed for those who want the utmost in sturdy construction and electrical flexibility in coils handling up to 500 watts. These famous coils pioneered individual adjustable link construction, thus providing precise impedance matching up to 600 ohms.

B & W COAXIAL CONNECTOR CC-50

Provides efficient, watertight coaxial cable connections for amateur and commercial use. Also serves as a center insulator. Made of aluminum with steatite insulation, this unit comes complete with weatherproof cement and assembly screws. Weight 12 oz., pull strength 500 lbs.



B & W HEAVY-DUTY VARIABLE CAPACITORS

Type CX is a radically designed split stator, butterfly rotor variable capacitor that permits mounting the tank coil assembly directly on the capacitor frame as illustrated. Opposed stator sections provide short R-F paths desirable in high power rigs. Built-in neutralizing capacitors are provided for, on rear end plate.

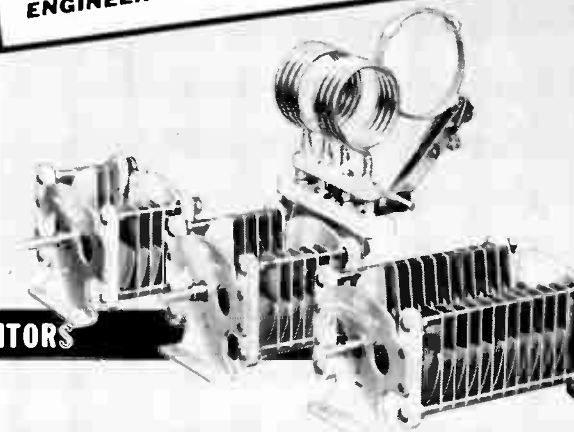
B & W PLUG-IN LINKS

For impedance matching, just plug in the proper link. These new B & W plug-in links make your rig adaptable to practically any impedance as quickly as you can plug in a link with the correct number of turns. Type 3750 mounting bar for HDV coils, Type 3550 for TVH-TVL-BVL coils. Links available in 1-3-6 and 10 turns.

BARKER & WILLIAMSON, Inc.

EQUIPMENT

PIONEERED, DESIGNED AND BUILT FOR EXACTING ELECTRONIC ENGINEERS AND EXPERIMENTERS



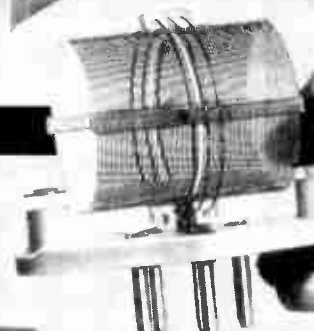
In addition to the B & W products shown here, there are dozens of others in our general catalog. All are made under the direct supervision of men who know amateur radio requirements personally. And all are produced to the high quality and design standards that are characteristic of B & W equipment.

B & W BUTTERFLY VARIABLE CAPACITORS

Type JCX small butterfly variable capacitors offer all the electrical and mechanical features of the larger B & W units, and are ideally small in size for general ham and other uses. They accommodate B & W "B" or "BX" series coils and offer many advantages over conventional type units.

B&W ALL-BAND FREQUENCY MULTIPLIER

Model 504—A fixed-tuned, broad-band frequency multiplier designed for use with either a V.F.O. or Crystal input. Makes transmission on any band available at the flip of a switch.



B & W "BABY" AIR INDUCTORS

25 Watts Rating—Ideal for crowded layouts, portables and any other application where space is at a premium and high efficiency a "must." Many other types and sizes available. All offer famous B & W "air wound" construction.

B & W TEST INSTRUMENTS

Accurate—Inexpensive—Reliable

AUDIO OSCILLATOR—*Model 200*—An extremely low distortion source of frequencies between 30 and 30,000 cycles.

DISTORTION METER—*Model 400*—Measures total harmonic distortion for the range of 50 to 15,000 cycles.

SINE WAVE CLIPPER—*Model 250*—Provides test signal particularly useful in examining the phase angle, transient and frequency response of audio circuits.

FREQUENCY METER—*Model 300*—An accurate and convenient means of making direct measurements of unknown audio frequencies up to 30,000 cycles.

WRITE FOR COMPLETE B & W INSTRUMENT CATALOG!

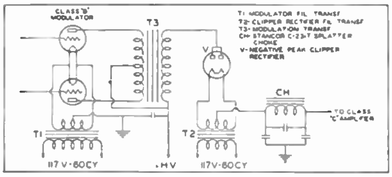
Get this Handy Catalog

for full details on inductors, variable capacitors and accessories every ham needs for his rig.

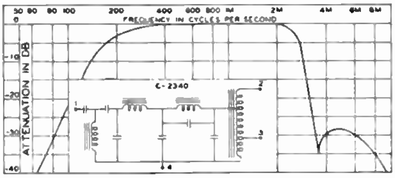
237 FAIRFIELD AVE., UPPER DARBY, PA.

Modernize YOUR PHONE TRANSMITTER with a STANCOR Audio Filter

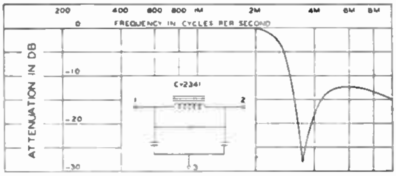
C-2317—Use of a splatter suppressor filter between the modulator and Class C amplifier eliminates undesirable high audio frequencies and harmonics which cause illegal interference to stations on other channels. Stancor Part Number C-2317, illustrated in a typical circuit application in the diagram at the right, attenuates frequencies higher than 3,000 CPS when used in accordance with supplied instruction data. The effectiveness of the system is greatly enhanced by the negative peak limiter shown in the circuit.



C-2340—In radiotelephony, it is highly desirable to limit frequencies in the side bands to those providing the greatest degree of intelligibility. Useless, power-consuming frequencies below 200 cycles and above 3,000 cycles can be efficiently eliminated by insertion of the Stancor C-2340 band-pass filter in the speech amplifier. When used in conjunction with a peak clipper, a high *average* percentage of modulation is possible, providing a signal that rides over the QRM. The graph illustrates the frequency curve of this three-section M-derived filter.



C-2341—Where the properties of a low-pass filter meet circuit requirements, the economical M-derived Stancor C-2341 will give a good account of itself and may be used to advantage with a peak clipper. Typical circuit application of this low-pass filter may be found on page 24 of the November, 1946, issue of *QST*. The frequency curve of the C-2341 is shown at the right.



Your electronic parts distributor can supply you with more than four hundred types of Stancor transformers and related components for radio, television, sound and other electronic applications. In addition, special units can be built to your specifications for experimental work.

Competent Stancor engineering, advanced manufacturing techniques and rigid standards of quality ensure components that improve the performance of the finest equipment.

This Helpful Stancor Literature Is Yours for the Asking

GENERAL CATALOG 400 transformers and related components are completely described in Stancor's general catalog. Also included are handy reference charts.

TELEVISION COMPONENTS Revised frequently to keep pace with the fast-growing television industry, Catalog 337 lists all Stancor TV components currently available, including the "Exact Duplicate" series for dependable replacement.

HIGH FIDELITY The outstanding HF and WF series of high fidelity transformers were designed to meet the demand of audio engineers and enthusiasts for superlative audio components. Stancor Catalog 336 gives complete transformer specifications, including frequency response data.

MOBILE TRANSMITTER The pleasures of mobile operation are being discovered by more and more amateurs every day and their popular transmitter choice is the field-proven ST-203-A unit, completely described, with illustrations, in Catalog Sheet 1311R.

PICK UP COPIES OF THIS STANCOR LITERATURE AT YOUR DISTRIBUTOR OR WRITE

STANDARD TRANSFORMER CORPORATION
3590 ELSTON AVENUE CHICAGO 18, ILLINOIS

THERE ARE MANY GRID-DIP METERS

But Only ONE

MEGACYCLE METER

Model 59

WITH THESE EXCLUSIVE FEATURES:

- WIDE FREQUENCY RANGE— 2.2 MC TO 400 MC.
- FREQUENCY CALIBRATION ACCURATE TO ± 2%.
- 120 CYCLE MODULATION.
- OSCILLATOR-PROBE SEPARATE FROM POWER SUPPLY FOR EASE OF OPERATION.
- AC OR BATTERY OPERATION.
- PRECISION INSTRUMENT DESIGN AND CONSTRUCTION.

Specifications:

FREQUENCY: 2.2 Mc. to 400 Mc.; seven plug-in coils.

MODULATION: CW or 120 cycles; or external.

OSCILLATOR UNIT: 3¾" diameter; 2" deep.

POWER UNIT: 5⅛" wide; 6⅛" high; 7½" deep.

POWER SUPPLY: 117 volts, 50-60 cycles; 20 watts.

A Multi-Purpose Instrument For Amateur, Service Man, Engineer

- For determining the resonant frequency of tuned circuits, antennas, transmission lines, by-pass condensers, chokes, coils.
- For measuring capacitance, inductance, Q, mutual inductance.
- For preliminary tracking and alignment of receivers.
- As an auxiliary signal generator; modulated or unmodulated.
- For antenna tuning and transmitter neutralizing, power off.
- For locating parasitic circuits and spurious resonances.
- As a low sensitivity receiver for signal tracing.

MANUFACTURERS OF
Standard Signal Generators
Pulse Generators
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UHF Radio Noise & Field Strength Meters
Capacitance Bridges
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Television and FM Test Equipment

TELEVISION INTERFERENCE

The Model 59 will enable you to make efficient traps and filters for the elimination of most TV interference.

Write for Free Data Sheet 59TVI

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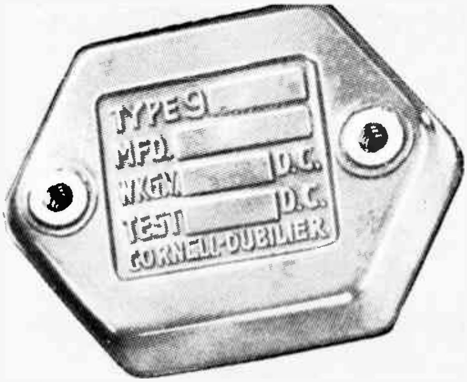
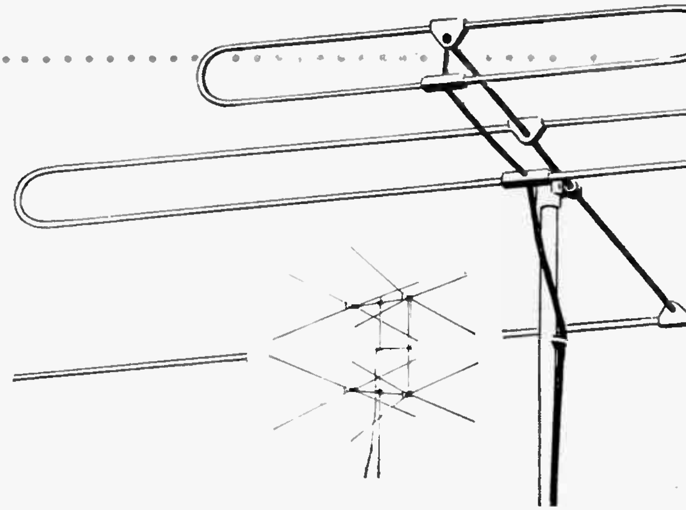


CORNELL-DUBILIER CAPACITORS

Cornell-Dubilier Paper, Mica, Dykanol and Electrolytic Capacitors for every conceivable application. For over forty years the capacitor line most in demand. Send for Catalog No. 200B for full description.

CORNELL-DUBILIER "SKYHAWK" ANTENNAS

Cornell-Dubilier TV, FM, AM and Auto Antennas in all shapes and sizes for every possible application. Send for descriptive literature.



CORNELL-DUBILIER VIBRATORS

Both heavy duty and automobile radio vibrators have now been added to the famous C-D bannerhead. For the finest engineering in vibrators – for quiet, stable long-life insist on C-D vibrators. Catalog No. VA of standard stock vibrators on request.

CORNELL-DUBILIER POWERCON CONVERTERS

C-D Powercons are honestly rated for dependable trouble-free long life. A complete line for conversion to 110 volts AC from 6, 32 or 110 volts DC; converters for operating phono-turntables from 110 VDC; battery chargers for 6 and 12 volt DC output from 110 VAC and converters for conversion of DC voltage at one level to DC at another. Powercon catalog on request.

THESE
ASTATIC MICROPHONES
Now Enjoy a
DOUBLE POPULARITY

THE ALL-TIME, leading popularity of Astatic Microphones now goes DOUBLE. All models shown are available with ceramic as well as crystal elements. The growing acceptance for the ceramic types has placed them almost shoulder to shoulder—in point of preference—with the tried-and-true favorites, the crystal units. Here, to aid you in your personal choice, is the technical data on each:

D-104 SERIES

JT SERIES

T-3 SERIES

VELVET VOICE SERIES

SPECIFICATIONS

Model	Output Level	Range	Response Characteristics
D-104	—48 db.	30-7,500	Rising
T-3	—52 db.	30-10,000	Substantially flat
JT-30	—52 db.	30-10,000	Substantially flat
JT-40	—52 db.	30-10,000	Rising
200	—52 db.	30-10,000	Substantially flat
241	—52 db.	30-10,000	Rising
D-104-C	—58 db.	30-7,500	Rising
T-3-C	—62 db.	30-10,000	Substantially flat
JT-30-C	—62 db.	30-10,000	Substantially flat
JT-40-C	—62 db.	30-10,000	Rising
VC	—62 db.	30-10,000	Substantially flat
VC-1	—62 db.	30-10,000	Rising

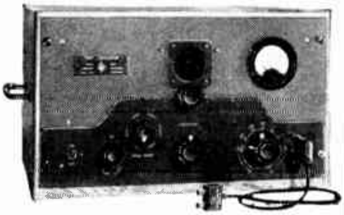
Letter "C" in model number designates ceramic unit.

THE Astatic CORPORATION
ASTATIC
CONNEAUT, OHIO
IN CANADA CANADIAN ASTATIC LTD TORONTO ONTARIO

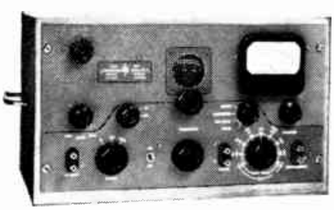
Astatic Crystal Devices manufactured under Brush Development Co. patents

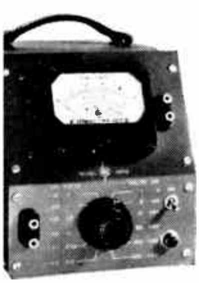
HEWLETT PACKARD


-hp- 200C Audio Oscillator


-hp- 650A Signal Generator


-hp- 616A
UHF Signal Generator

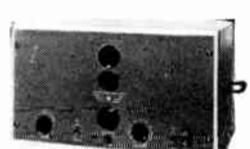

-hp- 330B Distortion Analyzer


-hp- 400C
acuum Tube Voltmeter

FUNCTION	MODEL	FREQUENCY	CHARACTERISTICS
HARDWARE	10		Binding Post
	14		Flexible coupler, ceramic insulated; permits misalignment of 1/32" and/or 5°
LOW FREQUENCY STANDARDS	100A	100 kc, 10 kc, 1 kc, 100 cps	Accuracy 3 cps per mc per degree Centigrade
	100B	100 kc, 10 kc, 1 kc, 100 cps	Temperature controlled; accuracy 0.001%
FREQUENCY DIVIDER	110	100 to 10 cps	Controlled by 100A or 100B. Multipliers also available up to 1 mc
RESISTANCE-TUNED OSCILLATORS	200A	35 to 35,000 cps	Output 1 watt into 500 ohms; 1% distortion
	200B	20 to 20,000 cps	Output 1 watt into 500 ohms; 1% distortion
	200C	20 to 200,000 cps	Output 10 volts into 1,000 ohms; 1% distortion
	200D	7 to 70,000 cps	Output 10 volts into 1,000 ohms; 1% distortion
	200H	60 to 600,000 cps	Output 10 mw into a 100 ohm load; 3% total distortion
	200 I	6 to 6,000 cps	Frequency setting closer than 1%; output 10 volts into 1,000 ohms; 1% distortion
	201B	20 to 20,000 cps	Output 3 watts at 1% and 1 watt at ½% distortion into 600 ohms
	202B	½ to 50,000 cps	For low frequency studies. Output 10 volts into 1,000 ohms; 1% distortion
	202D	2 to 70,000 cps	Output 10 volts into 1,000 ohms; 2% distortion
	204A	2 to 20,000 cps	Portable, battery-operated; output 5.0 volts to 10,000 ohm load; 1% distortion
AUDIO SIGNAL GENERATORS	205A	20 to 20,000 cps	Output 5 watts, 1% distortion into impedances of 50, 200, 600, 5,000 ohms. Output VTVM and 110 db attenuator, 1 db steps
	205AG	20 to 20,000 cps	Same as 205A, plus separate VTVM for complete gain measurements
	205AH	1 to 100 kc	Output 5 watts, 3% distortion into 50, 200, 500, 5,000 ohm impedances. Output VTVM and 110 db attenuator, 1 db steps
	206A	20 to 20,000 cps	Output +15 dbm with less than 0.1% distortion into 50, 150, 600 ohm impedances. Output VTVM and 111 db attenuator in 0.1 db steps
SQUARE WAVE GENERATOR	210A	20 to 10,000 cps	Output 50 volts peak to peak; 1,000 ohm internal impedance; 70 db attenuator, 5 db steps
WAVE ANALYZER	300A	30 to 16,000 cps	Variable selectivity; measurement range 1 mv to 500 volts; 5% accuracy

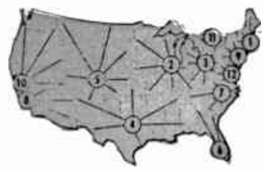

-hp- 410A Vacuum Tube Voltmeter


-hp- 805A Slotted Line


-hp- 202B Low Frequency Oscillator

HEWLETT -hp- PACKARD

ANOTHER -*hp*- SERVICE



Person-to-Person Attention to Your Measuring Problems

Almost anywhere in America, -*hp*- field representatives can provide personal help with your measuring problems. They have complete data on all -*hp*- instruments—their performance, servicing, and adaptabilities.

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2017 S Street N.W.
Decatur 8000

For details on any -hp- instrument, contact the nearest -hp- representative or write direct to factory.

FUNCTION	MODEL	FREQUENCY	CHARACTERISTICS
DISTORTION ANALYZERS	320A	400 cps and 5 kc	Measures total distortion as low as 0.1%. 70 db attenuator, 1 db steps for comparison
	320B	50, 100, 400 cps and 1, 5 and 7.5 kc	Same as 320A
	325B	30, 50, 100, 400, 1,000 cps; 5, 7.5, 10 and 15 kc	Measures total distortion as low as 0.1%. Input amplifier and complete VTVM each usable separately
	330B	Any frequency 20 to 20,000 cps	Similar to 325B but measures at any frequency and includes AM detector
	330C	Any frequency 20 to 20,000 cps	Similar to 330B, no AM detector. Meter has VU characteristics to meet FCC requirements for FM broadcasting
FM BROADCAST MONITOR	335B	88 to 108 mc	FCC approved. Monitors carrier frequency and modulation. High fidelity output for aural monitoring
ATTENUATORS	350A	Max 100 kc	110 db, 1 db steps; 5 watts, 500 ohm level. Bridged T type. Accuracy 1 db in 50 db at 100 kc
	350B	Max 100 kc	Same as 350B but 600 ohm level
VACUUM TUBE VOLTMETERS AND ACCESSORIES	400A	10 cps to 1 mc	Nine ranges 0.03 to 300 volts full scale. Accuracy ±3% to 100 kc, ±5% to 1 mc. Average reading. Calibrated in rms.
	400B	2 cps to 100 kc	Same as 400A with response flat to 2 cps. 10 megohm input impedance
	400C	20 cps to 2 mc	Twelve ranges 0.001 to 300.0 volts full scale; accuracy ±3% to 100 kc, ±5% to 2 mc; 10 megohn input impedance; average reading; calibrated in rms volts; may be used as 54 db amplifier
	404A	2 to 50,000 cps	Portable, battery-operated; eleven ranges; 0.003 to 300 volts full scale; accuracy ±3% to 20 kc; 10 megohm input impedance
	410A	20 cps to 700 mc	AC: six ranges 1 to 300 volts. DC: seven ranges 1 to 1,000 volts. Resistance: seven ranges 0.2 ohm to 500 megohms
	415A	300 to 2,000 cps	Standing Wave Indicator for use with a bolometer or crystal rectifier; standard frequency 1000 cps, others on special order
	430A		Microwave Power Meter. Measures directly power dissipated in a standard barretter. Full scale reading 0.1 to 10.0 milliwatts.
	455A	to 1,000 mc	Connects probe of 410A across 50 ohm transmission line. Type N fittings
	458A	to 1,000 mc	Connects probe of 410A to open end of 50 ohm transmission line. Type N fittings
AMPLIFIERS	450A	10 to 1,000,000 cps	40 db and 20 db stabilized gain. Input impedance 1 megohm shunted by approximately 15 uuf.
ELECTRONIC FREQUENCY METER	500A	5 cps to 50 kc	Ten ranges, ±2% accuracy. Input 0.5 to 200 volts
ELECTRONIC TACHOMETER	505A	300 to 3,000 000 rpm	Ten ranges, ±2% accuracy
	505B	5 to 50,000 rps	Same as 505A except calibrated in rps
SIGNAL GENERATORS	610A	500 to 1,350 mc	Calibrated output 0.1 microvolt to 0.1 volt. Internal pulse modulation. Direct calibration
	616A	1,800 to 4,000 mc	Direct reading. Pulse modulation, CW and FM. Calibrated output 0.1 microvolt to 0.2 volts
	650A	10 cps to 10 mc	Direct reading. Six bands. Output 3 volts to 600 ohm load. VTVM and output attenuator
POWER SUPPLY	710A		Any dc voltage 180 to 360 for 0 to 75 ma load; approximately 1% regulation. Also 6.3 volts, 5 amps ac.
SLOTTED LINE	805A	500 to 4,000 mc	VSWR Less than 1.04. Tunable probe. 50 ohm impedance. Type N connectors.

Accurate measurements are fundamental to electronic progress. -*hp*- precision instruments are basic tools for making these measurements swiftly, surely, easily. -*hp*- instruments are used in laboratories, factories, radio and television stations, military installations and technical colleges throughout the world.

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FOR SPEED AND ACCURACY

MAKE THE MOST OF YOUR *Station Power*

VOLT-OHM-MILLIAMMETER MODEL 666-HH

A precision-made marvel of compactness that provides a complete miniature laboratory for voltage resistance, and direct current analyses.

RANGES

D.C.Volts: 0–10–50–250–1000–5000 at 1000 ohms-Volt.
A.C. Volts: 0–10–50–250–1000–5000 at 1000 ohms-Volt.
D.C. Milliamperes: 0–10–100–500 at 250 Millivolts.
OHMS: 0–2000–400,000.

NEW MODEL 666-R—A new pocket tester with precision resistors, unit construction and higher resistance ranges.

★ ★ ★

A COMPLETE LINE OF METERS

Triplett panel and portable meters are available in more than 26 case styles—round, square, and fan—2" to 7" sizes. Included are voltmeters, ammeters, milliammeters, millivoltmeters, microammeters, thermoammeters, DB meters, VU meters and electrod,namometer type instruments.

. . . WITH MODEL 3296

With Model 3296 you can modulate to the fullest, for top power output, yet you know at once when you are **over**-modulating. Four separate circuits measure amplitude modulation—(1) percent modulation, average; (2) peak flash percent modulation; (3) carrier shift; and (4) audio output for headphones. These may be used separately, all at once, or in any combination. Peak indicator can be pre-set for any percent of modulation from 20 to 120, to flash when pre-determined modulation is reached.

★ ★ ★

ABSORPTION FREQUENCY METER • Model 3256

A band-switching, tuned absorption type frequency meter that covers five amateur bands. Has new germanian crystal and a DC Milliammeter indicator for greater sensitivity. Direct calibration on panel—no coils to change. Switching permits instantaneous band change. Audio jack provides for monitoring of phone signals—another new feature. Fully shielded. Calibration is in megacycles in following bands: 3.5–4 MC; 7–7.3 MC; 14–14.4 MC; 20–21.5 MC; 28 30 MC.

Address all inquiries to Dept. AHB50

Triplett ELECTRICAL INSTRUMENT COMPANY
BLUFFTON, OHIO


COMPONENTS FOR EVERY APPLICATION
LINEAR STANDARD
High Fidelity Ideal
HIPERM ALLOY
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Portable . . . High Fidelity
OUNCER
Wide Range . . . 1 ounce
SUB OUNCER
Weight 1/3 ounce
COMMERCIAL GRADE
Industrial Dependability
SPECIAL SERIES
Quality for the "Ham"
POWER COMPONENTS
Rugged . . . Dependable
VARITRAN
Voltage Adjustors
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One watt to 100KW
VARIABLE INDUCTOR
Adjust like a Trimmer
TOROID HIGH Q COILS
Accuracy . . . Stability
TOROID FILTERS
Any type to 300KC
MU-CORE FILTERS
Any type 1/2 – 10,000 cyc.
EQUALIZERS
Broadcast & Sound
PULSE TRANSFORMERS
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SATURABLE REACTORS
Power or Phase Control
PLUG IN ADAPTER
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CABLE TYPE
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Match any line voltage
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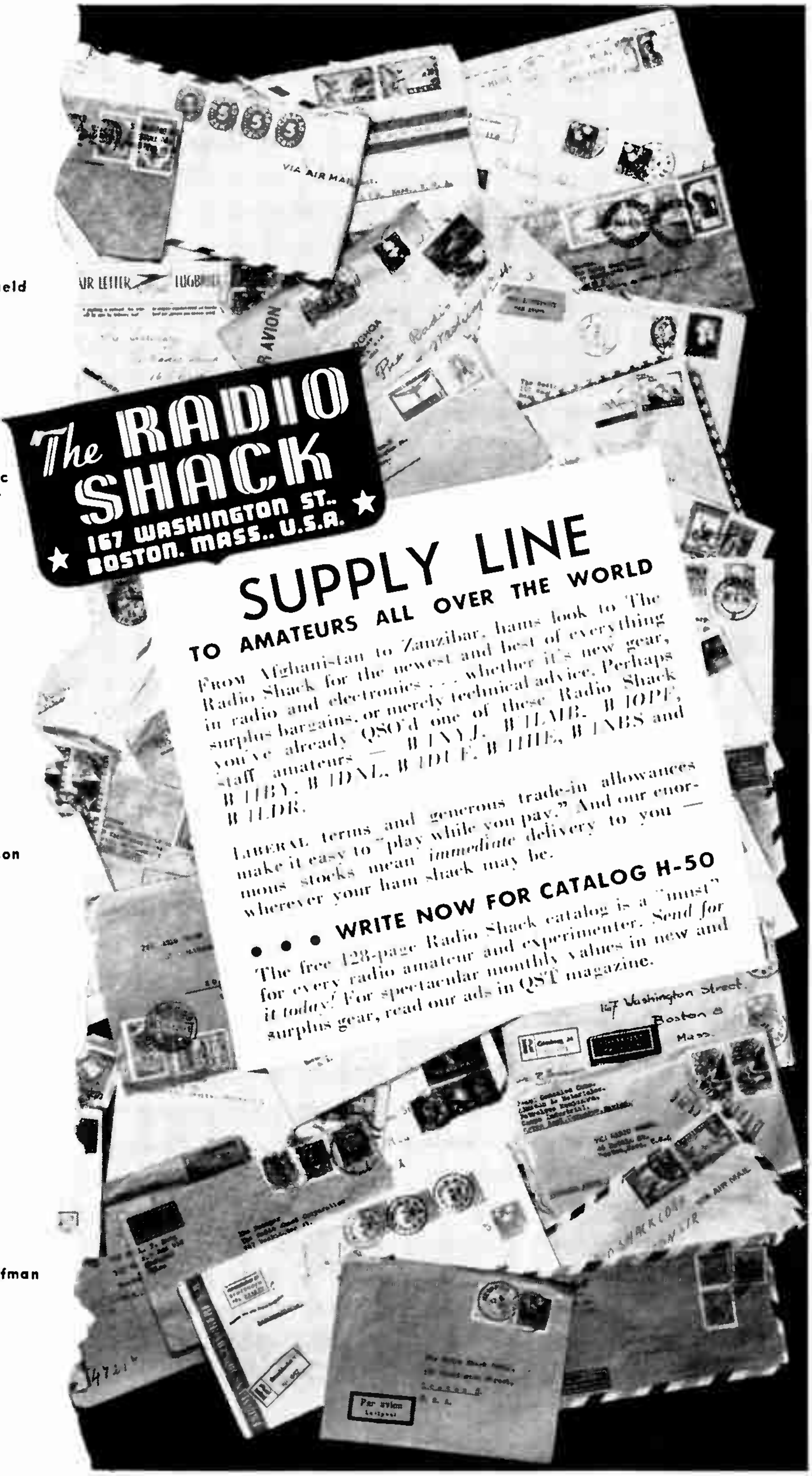

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TRANSMISSION LINE CABLES
Every Type for Every Service
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LOW PRICES: I sell to you as cheap or cheaper than you can buy anywhere.

COMPLETE STOCKS: Collins, Hallicrafters, National, Hammarlund, RME, Millen, Harvey-Wells, Meissner, Gonset, Meck, Johnson, RCA, all other amateur receivers, transmitters, beams, TV, AM-FM, high fidelity amplifiers and speakers, test equipment, tubes, parts, etc. I can supply nearly any equipment shown in any catalog or advertisement and at lowest prices.

BEST TRADE-IN ALLOWANCE: Customers in all parts of the USA trade with me because I allow so much. Tell me what you have to trade and what you want. I also buy equipment.

TIME PAYMENTS: You can order anything on terms. I finance the terms myself to save you time and bother. Customers everywhere in the USA find my terms best. Write for details.

QUICK DELIVERY: Mail, phone, or wire your order. It will be shipped promptly. I can be reached nearly 24 hours a day, 7 days a week.

TEN-DAY TRIAL: Try any communications receiver ten days -- if you return it your only cost is shipping charges.

PERSONAL ATTENTION: The Butler store is run by Bob Henry, WøARA, and the Los Angeles store by Ted Henry, W6UOU. We make the deals ourselves. We finance the time payments ourselves. That way we have the lowest overhead and can do more for you. That's why YOU AND I CAN DO BUSINESS. Write, phone, or visit either store.

73,

Bob Henry
WØARA

Butler, Missouri

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11240 Olympic Blvd.
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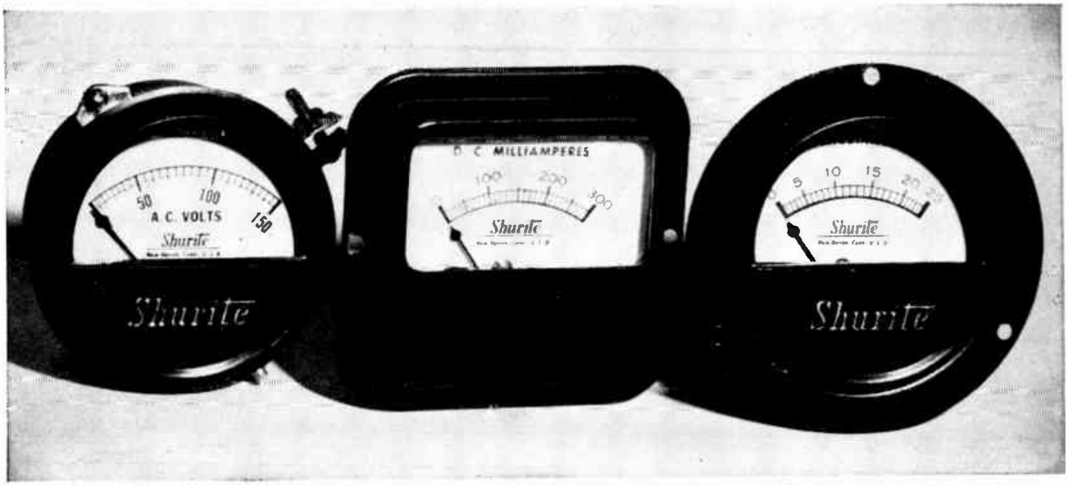
"WORLD'S LARGEST DISTRIBUTORS OF SHORT WAVE RECEIVERS"



MODEL 550-AC MODEL 950-DC MODEL 650-DC

Models shown are ⅔ actual size

Shurite PANEL METERS Meet Your Needs *because . . .*

They're RUGGED . . . . . . Sturdy construction throughout. Molded inner unit with coil frames and insulators integral for maximum rigidity. Exceptionally high ratio, torque-to-weight.

They're NEAT . . . . Dials are metal so they stay good looking in spite of age and moisture. Rich telephone black finish on metal cases. Concealed coils and good readable scales.

They're SENSITIVE . . . . . Accuracy well within 5%. AC meters are double-vane repulsion type. DC meters are polarized-vane solenoid type. High internal resistance models available in popular ranges.

They're GUARANTEED . . . For one year from date of purchase against defective workmanship and material, and will be repaired or replaced if sent to the factory postpaid with 25c handling charge.

They're INEXPENSIVE . . . . For instance, Model 950, 0–100 DC Ma. sells for $1.45; Model 550, 0–15 DC Amps. for $1.30. Other meters are correspondingly reasonable in price.

They're AVAILABLE . . . . Stocked by leading electronic distributors in a wide variety of types and ranges.

The line is COMPLETE . . . All of these features are available in 216 ranges and types: AC, DC; Voltmeters, Ammeters, Milliammeters, Resistance Meters. For instance, DC Milliammeters are made in 60 types and ranges.

Ask for General Catalog Sheet F53-SH

Shurite also makes pocket meters and testers. Ask your distributor for further information, or write us direct

SHURITE METERS

87 HAMILTON STREET, NEW HAVEN 8, CONN.

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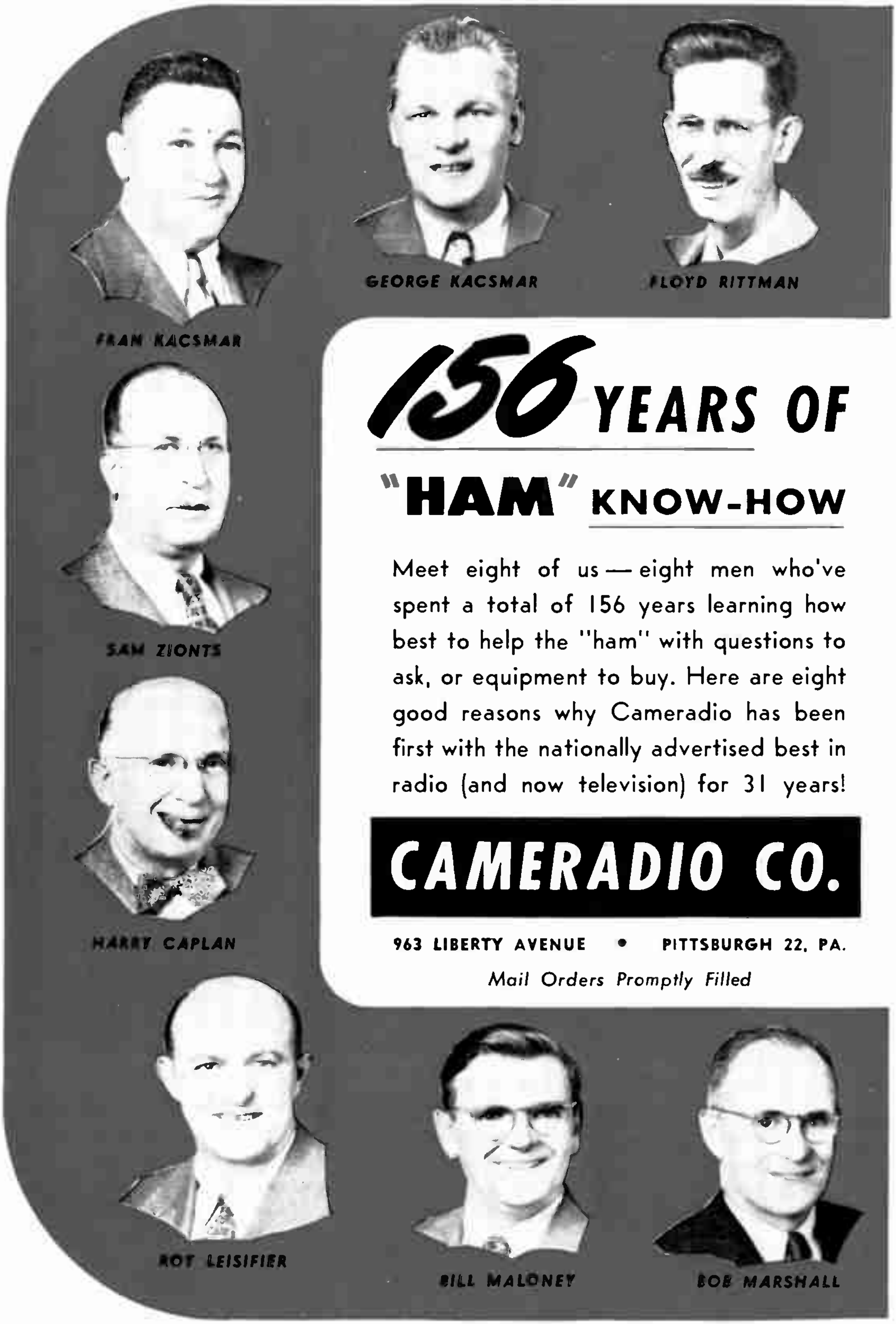
TRADE MARK

ELECTRICAL MEASURING INSTRUMENTS



FRAN KACSMAR
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156 YEARS OF
"HAM" KNOW-HOW
Meet eight of us — eight men who've spent a total of 156 years learning how best to help the "ham" with questions to ask, or equipment to buy. Here are eight good reasons why Cameradio has been first with the nationally advertised best in radio (and now television) for 31 years!
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963 LIBERTY AVENUE • PITTSBURGH 22, PA.
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"Why don't you TELL THE HAM?"

"Why hide your light under a bushel? Tell the ham: In commercial circles Hytron is even more famous for fine *receiving* tubes. Any ham knows Hytron's reputation for dependable transmitting and special purpose tubes: The popular 5514 triode. Vhf favorites; HY75A, HY615. Instant-heating 2E25A, 2E30, HY69, 5516, 5812, etc. Top-grade voltage regulators; OD3/VR150, OB2, etc.

"But does he know these facts? Hytron is the oldest manufacturer specializing in receiving tubes. Hytron originated the subminiature . . . the GT . . . over 50 GT types . . . many new miniatures . . . a new line of low-cost receiving tubes for TV. Hytron works closely . . . constantly with leading set manufacturers to improve already superior receiving tubes. Hytron offers the ham premium-quality receiving tubes for the tougher TV jobs—at no extra cost. And Hytron makes doggone fine TV Picture Tubes.

"Does the ham know Hytron has for him all kinds of receiving tubes . . . GT, G, standard glass, metal, lock-in, 7-pin and 9-pin miniatures? Heck, *tell* the ham. Hytron is his best bet. For dependable receiving tubes—as well as transmitting and special purpose tubes."

Thus did one of our best friends tell us! Now he has helped us tell you. You know too!

"YOU CAN'T BEAT HYTRON FOR RECEIVING TUBES"

NEW HYTRON SHOP TOOLS

Economical. Original. Practical. Time-and-effort savers. Soldering Aid . . . Tube Lifter . . . Tube Tapper . . . Miniature Pin Straighteners, etc. Designed by and for radio servicemen in recent Hytron Contest. Ideally suited to ham use too. At Hytron jobbers. Write today for free Hytron Shop Tool Catalogue.

FREE

1. Retail Price List for Hytron Receiving Tubes.
2. Hytron Reference Guide for Miniature Electron Tubes, 3rd Ed.
3. Hytron Transmitting and Special Purpose Tube Catalogue.
4. Complete data sheets: 2E25A, 2E30, 3B4, HY31Z, HY69, HY75A, HY1231Z, HY1269, 5514, 5516, 5812.

OLDEST MANUFACTURER SPECIALIZING IN RECEIVING TUBES

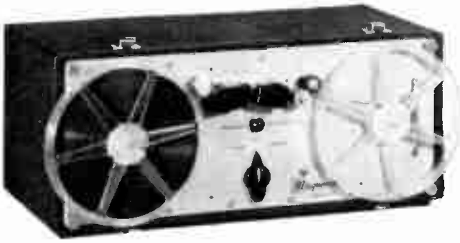
HYTRON

RADIO AND ELECTRONICS CORP.

MAIN OFFICE: SALEM, MASSACHUSETTS

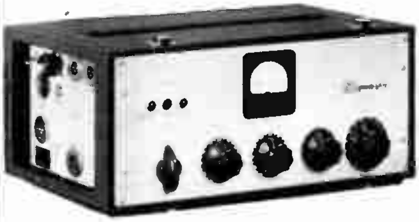
HARVEY TWO GREAT NAMES MAGNECORD

Combine these units to suit your needs and your purse. For portable or studio use. Conforms to all N.A.B. specifications. Precision capstan and drive.



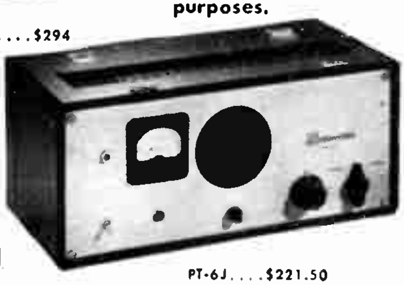
PT-6A . . . $278 and PT-6AH . . . $294

The PT-6A Tape Recorders are now in use in hundreds of broadcast stations and recording studios. Two tape speeds: 7½ and 15 inches plus high speed forward on the PT-6AH for cueing purposes.



PT-6P . . . $462

BROADCAST QUALITY RECORDING AND REPRODUCTION



PT-6J . . . $221.50

Designed for professional use, these units provide the highest quality. Frequency response ± 2 db from 40 to 15000 cycles, less than 2% harmonic distortion at full modulation. The recorder has high speed rewind (45 seconds) and high speed forward is available on the PT-6AH at slightly higher cost. The PT-6P amplifier is a fully portable record-playback-remote unit with 3 mike inputs and monitor speaker. The PT-6J has single mike input monitor speaker, jack for external speaker, can be used for public address. Full details of these remarkable units require pages. Come in for a demonstration or write us for full details.

PT-6A	Basic Recorder mechanism, includes 7½ and 15 inch capstans, interconnecting cords and portable carrying case	$278.00
PT-6AH	Basic recorder, same as above, with high speed forward	294.00
PT-6P	Portable Mixer Amplifier, in portable case	462.00
PT-6J	Amplifier	221.50
PT-6R	Rack mount amplifier	383.00
PT-6H	Rack Panel, for mounting PT-6A	7.00
PT-6M	Auxiliary spooling mechanism	128.00
PT-6T	Throwover Switch for using 2 PT-6A units with 1 PT-6P amplifier	32.00

...and the MAGNECORD PT-6R RACK AMPLIFIER

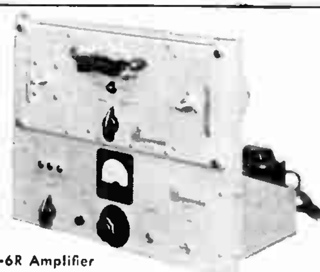
For studio use, or for interchangeable studio and remote, the PT-6R provides the studio half of the combination. Designed to take the PT-6A or PT-6AH recording mechanism (*as shown in illustration*), the recording unit may be lifted out in a few seconds and placed in its case for portable use.

This amplifier combines stable, broadcast-quality operation with solid construction. It has a single input channel and gain control providing either 600 ohms zero level, or high impedance bridging input. Pre- and post-emphasis equalization is built-in, compensating for tape magnetic characteristics. Program is fed to the PT-6R flat, and the playback is fed to the line flat. 600-ohm line output; signal develops approximately 1 ma in the PT-6A recording head. Response is flat, ± 2 db, 40 cps to 15 kc at 15" tape speed; 40 cps to 7 kc at 7½" tape speed. Three-position switch selects "record" or "listen", proper equalization being inserted for either operation, third switch position removes all equalization.

Volume level meter is supplemented by headphone jack on front panel. Uses 1-12AX7, 2-12AU7, 1-6X4, and supplies operating power for the bias oscillator unit in the PT-6A recording mechanism. 117 volts, 60 cycle, 1 phase, 60 watts. Standard 19" relay rack panel, 14" high, 12½" deep.

Complete, including tubes . . . $383.00

All in stock for immediate delivery.



PT-6R Amplifier
with PT-6A Recorder Mounted

All prices Net, F.O.B. New York and subject to change without notice.

Telephone: LUxemburg 2-1500

hrc

HARVEY RADIO COMPANY INC.

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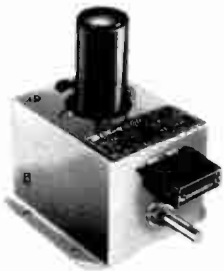
Bliley CRYSTALS

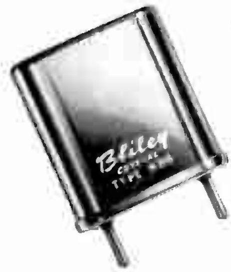
Experienced amateurs know Bliley craftsmanship!



TYPE CCO-A OSCILLATOR WITH AX2 OR AX3 CRYSTAL. Amateurs depend upon Bliley for high activity, precision calibrated, plated crystals. Supplied for 20-40-80-160 meters. Complete oscillators specifically designed for crystal control on 2-6-10-11 meters.

Bliley CRYSTALS

Bliley crystals are recommended when the end use is — Military!



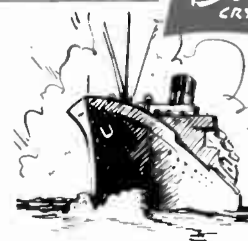
TYPE BH6. Commercial designation for HC-6 U holder; metal-to-glass hermetic seal. Supplied as type CR-18, CR-19, CR-23, etc. To Army, Navy, Air Force and CAA Specifications.

Bliley CRYSTALS

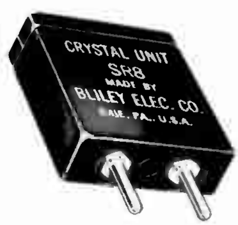
For mobile applications — Bliley crystals are usually specified!

TYPE MC7 CRYSTAL. Supplied as plated crystal in range 800-1800 kc. This new assembly eliminates microphonics and sets new precision standards for FM (30-50MC) equipment.

Bliley CRYSTALS



Aboard ship . . . Bliley crystals insure vital communications channels!



TYPE SR8 CRYSTAL. A rugged dependable frequency control recommended for ship-to-shore radiotelephone service. Gasket sealed case affords complete protection under adverse service conditions.

Bliley CRYSTALS

In the air . . . Bliley crystals help pin-point the flight!

TYPE SR5 CRYSTAL. Supplied with CAATC, when specified. This unit is commercial equivalent of military type CR-1A unit. Supplied as original equipment in many communications systems.

Bliley CRYSTALS

Call on Bliley for these special applications:

1. *ULTRASONICS:* Quartz blanks produced to customer specifications for ultrasonic application or research. Material carefully selected for freedom from flaws and processed to individual requirements.
2. *DELAY LINES:* Fused quartz delay lines custom-built to specifications.
3. *CRYSTAL OSCILLATORS:* Design, engineering, and production facilities for crystal oscillator subassemblies used in VHF equipment.

A complete knowledge of frequency control based on two decades of experience.

Bliley CRYSTALS

BLILEY ELECTRIC COMPANY

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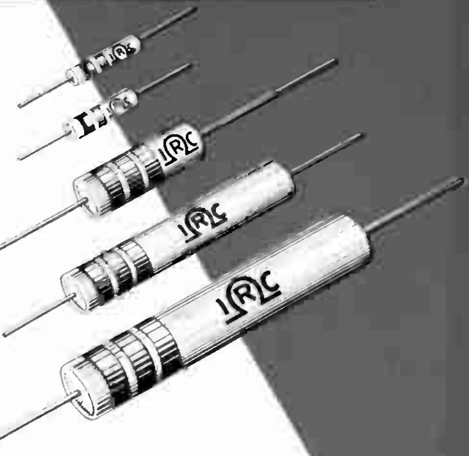


ADVANCED TYPE BT RESISTORS

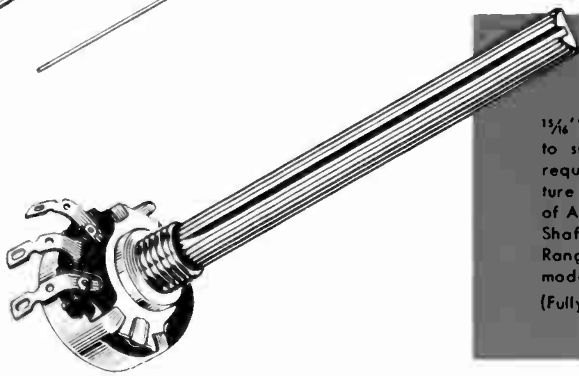
New type BT insulated Composition Resistors—meet JAN-R-11 Specifications at ⅓, ½, 1 and 2 watts. Small size BTB specially designed for miniature 2 watt requirements. Type BT's are suited to television and similar exacting circuits. Extremely low operating temperature. Excellent power dissipation. 330 ohms to 22 megohms in RMA ranges.

(Fully described in Catalog RDC8.)

BW INSULATED WIRE WOUND RESISTORS

Exceptionally stable, inexpensive low wattage wire wound resistors. ½, 1 and 2 watts—0.24 ohms to 8,200 ohms in RMA ranges. 50% to 100% overloads can be applied with negligible change, and return to initial value.

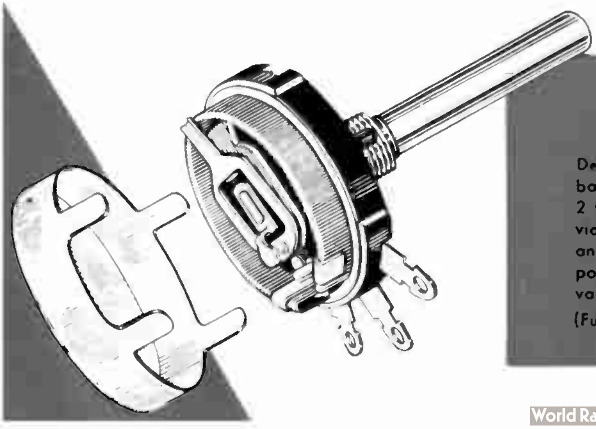
(Fully described in Catalog RDC8.)



NEWLY DEVELOPED TYPE Q VOLUME CONTROL

$1\frac{5}{16}$'' diameter and ¼'' bushing suit Type Q's to smallest chassis, yet they handle big-set requirements. Interchangeable *Fixed* Shaft feature (13 special shafts) gives coverage of 90% of AM, FM and TV needs. Knob Master *Fixed* Shaft fits most push-on knobs without alteration. Range: 500 ohms to 10 megohms. Accommodate Type '76 Switch.

(Fully described in Catalog RDC1-A.)



2 WATT RHEOSTAT-POTENTIOMETER

Designed for long, dependable service and balanced performance in every characteristic. 2 watt, variable wire-wound W Controls provide maximum adaptability to most rheostat and potentiometer applications within their power rating. Size 1¼'' by $\frac{9}{16}$''. Resistance values: 2 ohms to 10,000 ohms.

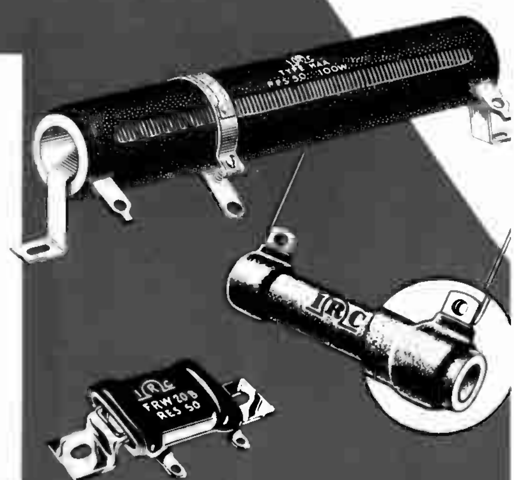
(Fully described in Catalog RDC1-A.)

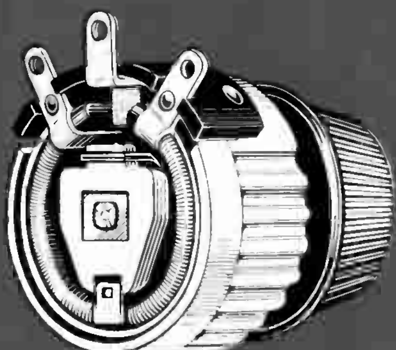
POWER WIRE WOUND RESISTORS

Fixed and adjustable Power Wire Wounds—10 to 200 watts—handle full rated power in all standard ranges, require no derating at high ranges. Dark, rough coating dissipates heat more rapidly. Unique terminals assure easy installation. 10 and 20 watt fixed types have lead and lug terminal, and lug may be clipped off for space saving in crowded chassis. Permanent, *fadeless* marking shows type, size, resistance.

Where limited space is a factor, Type FRW Flat Wire Wounds give higher space-power ratio than standard tubular types. Construction allows easy vertical or horizontal mounting, singly or in stacks.

(Fully described in Catalogs RDC-5 and RC-1.)

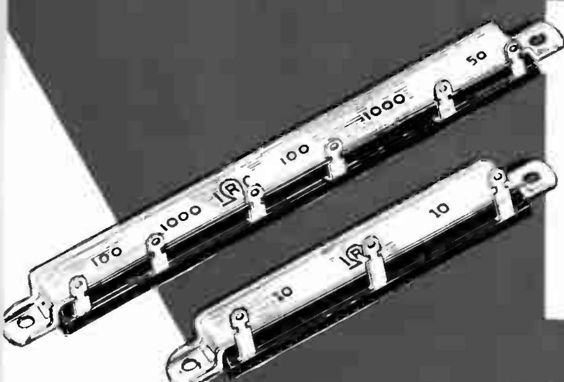




POWER RHEOSTATS

For variable power applications, IRC Type PR 25 and 50 watt Rheostats offer many advantages over conventional types. Operating temperatures are cut almost in half by aluminum construction. They can be used at full power in as low as 25% of rotation without appreciable temperature rise.

(Fully described in Catalog RDC-5.)



FLAT INSULATED WIRE WOUND RESISTORS

Unsurpassed for adaptability to an extremely wide variety of design requirements. Radical design features impervious phenolic compound casing, special metal mounting bracket that actually speeds transfer of heat from inside chassis. Space-saving MW's afford unusual flexibility in providing taps for voltage dividing applications.

(Fully described in Catalog RB-2.)

PRECISION WIRE WOUND RESISTORS

Combine the maximum in accuracy and dependability. Widely used in precision test equipment. 1% accuracy is standard; closer tolerances available at slightly increased cost.

(Completely described in Catalog RDC-6.)

Other Products in IRC's complete resistor line are described on the following pages.

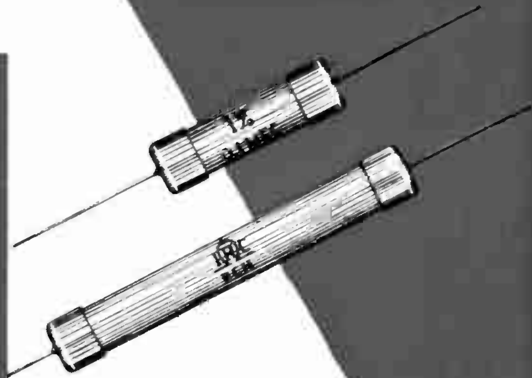
INTERNATIONAL RESISTANCE COMPANY

401 N. Broad Street Philadelphia 8, Pa.

In Canada: International Resistance Co., Ltd., Toronto, Licensee

IRC

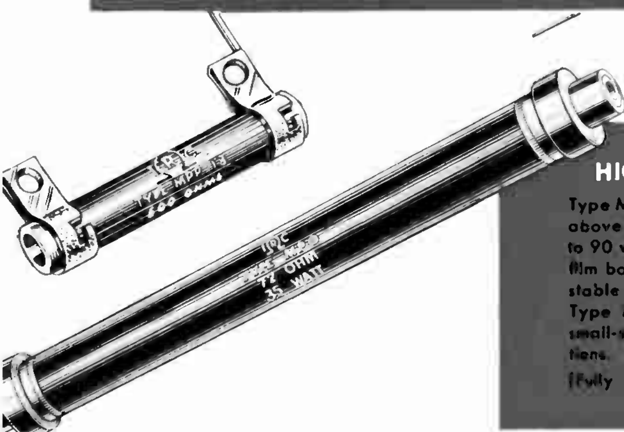
WHEREVER THE CIRCUIT SAYS



CLOSE TOLERANCE DEPOSITED CARBON PRECISTORS

PRECISTORS offer a unique combination of close tolerance, stability and economy. Pure crystalline carbon bonded to selected ceramic cores overcomes limitations of carbon composition resistors and higher cost of precision wire wounds. PRECISTORS offer wide range of values, guaranteed accuracy, high stability, low voltage coefficient, excellent frequency characteristics, predictable temperature coefficient.

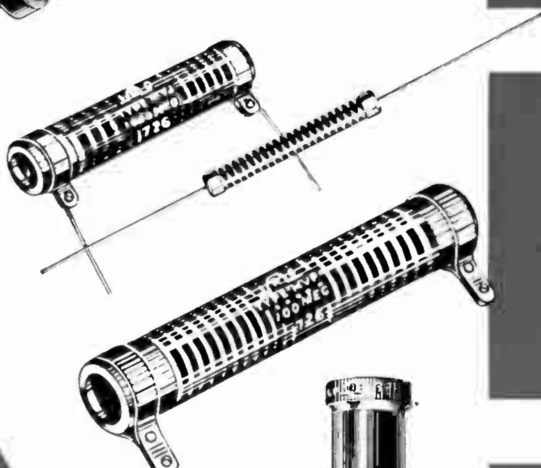
(Fully described in Catalog RDC-2.)



HIGH FREQUENCY RESISTORS

Type MP Resistors are designed for frequencies above those of conventional resistors. 2 watts to 90 watts. Special construction, with resistance film bonded to steatite ceramic form, provides stable resistors of low inductance and capacity. Type MPM's are miniature ¼ watt units for small-space, high frequency receiver applications.

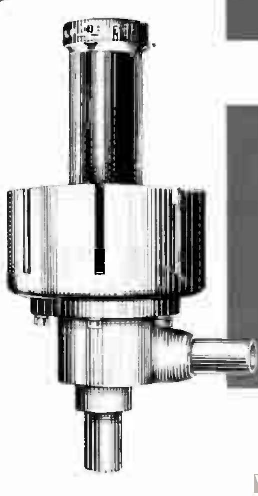
(Fully described in Catalog RF-1.)



HIGH VOLTAGE RESISTORS

Type MV's meet high resistance and power requirements in high voltage applications. Resistance coating in helical turns on ceramic tube provides a conducting path of long effective length. 2 watts to 90 watts. Variety of terminal types. Type MVX's meet requirements for small, high range unit with axial leads. 2" x ¼" construction identical with Type MV's, except for terminal.

(Fully described in Catalogs RG-1 and RG-2.)



WATER COOLED RESISTORS

Unique high frequency—high power resistor for television, FM and dielectric heating applications. Centrifugal force whirls high velocity stream of water in spiral path against resistance film—gives efficient high power dissipation up to 5 K.W. 35 ohms to 1,500 ohms. Resistor elements interchangeable.

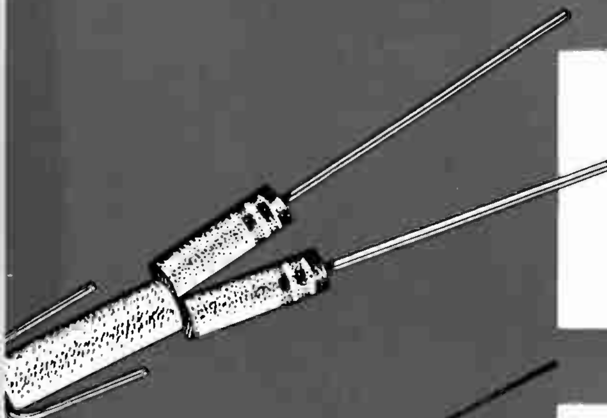
(Fully described in Catalog RF-2.)

Other products in IRC's complete resistor line are described on the preceding pages.

SEALED VOLTMETER MULTIPLIERS

Dependable multipliers for use under the most severe humidity conditions, Type MF Resistors consist of a number of IRC Precisions interconnected and hermetically sealed in a glazed ceramic tube. Compact, rugged, stable, fully moisture-proof and easy to install. Maximum current: 1.0 M.A.; 0.5 megohms to 6 megohms.

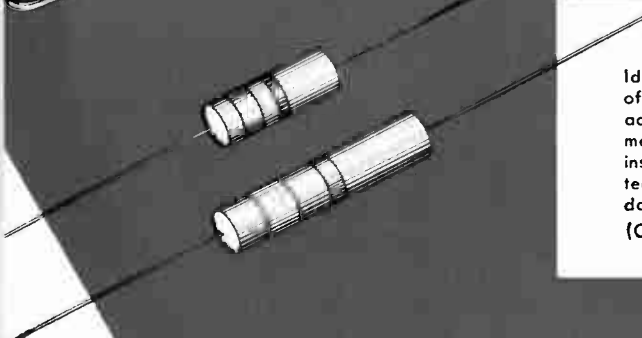
(Fully described in Catalog RD-2.)



MATCHED PAIR RESISTORS

Two resistors matched in series or parallel to as close as 1% initial accuracy. Dependable low-cost solution to close tolerance requirements. Both Types BT and BW resistors are available in matched pairs. Tolerances from ±5% to ±1% can be furnished.

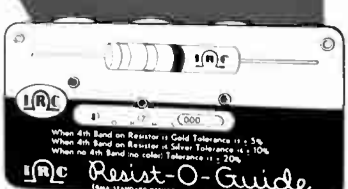
(Fully described in Catalog RB-3.)



INSULATED CHOKES

Ideal for TV and similar circuits. Wide range of size and characteristic combinations permit accurate specification to individual requirements. Types CLA and CL-1 Chokes are fully insulated in molded phenolic housings—protected from high humidity, abrasion, physical damage or shorting to chassis.

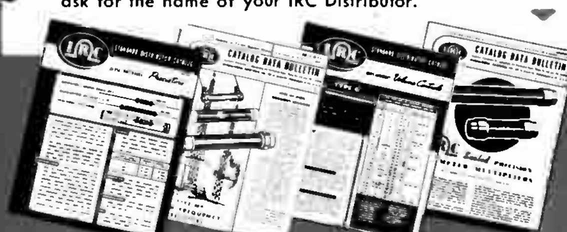
(Completely described in Catalog RDC7.)



IRC RESIST-O-GUIDE

New aid in easy resistor range identification. Turn 3 wheels to correspond with color code on resistors and standard RMA Range is automatically indicated. 15c at all IRC Distributors. When ordering direct, send stamps or coin.

For full information on any of IRC's many resistor types, write today for catalog bulletins in which you are interested. Also, ask for the name of your IRC Distributor.



INTERNATIONAL RESISTANCE COMPANY

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ESICO

REG. U. S. PAT. OFF.

• Red Label Irons

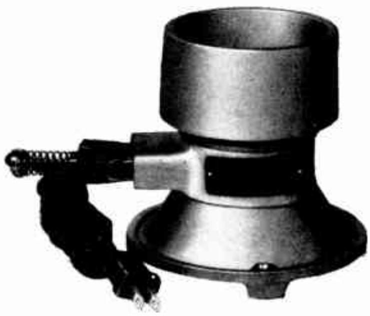
These are the irons that are used so universally in factory production lines. They are light weight, finely balanced, and have the coolest handles of any irons on the market. Elements are mounted and held in place with a knurled nut which engages the back end of the element and seats against the shoulder of the case shell, holding the element firmly in place regardless of the most rugged use. They are ideal from a maintenance standpoint for, due to their two piece combination terminal and handle, elements are replaceable in three minutes or less. The only iron on the market designed for use with or without a ground wire.

Irons are normally supplied in four wattages. They are obtainable, when required in quantity, in special wattages at no extra cost. Standard voltages are 105–120 and 220–240. Special voltages may be had. List prices of irons are as follows: **No. 38**—100 watt **$6.95, No. 58**—200 watt **$8.95, No. 78**—300 watt **$10.95** and **No. 98**—550 watt **$12.95.** The iron illustrated is the No. 38 and is ⅓ actual size.

• No. 61 Pencil Iron

This pencil iron is only seven inches in length and weighs just 2½ ounces exclusive of cord. The handle temperature at the point where it is held in the fingers, is actually no higher than body temperature. Diameter of handle is ¾″ and may be used as a pencil for the most delicate soldering operations. The element construction is of the same type as used in ESICO industrial irons and will give long service. The tip is the so-called plug type, held in place with a set screw. Three shapes of tips are available, Type B—¼″ dia. pyramid point, Type A—⅛″ dia. straight pencil point and Type C—⅛″ dia. bent 90 degrees with a pencil point.

The No. 61 is regularly wound to 25 watts at 105–120 volts, but may be had in higher wattages, when required in quantities at no extra cost. List price of iron is $4.95. Tips A, B, or C 30c each list. Irons are available thru any of the better tool or Radio & Electronic Supply houses. If your distributor can not supply you from stock, send your order direct to us, but please be sure and give name of your distributor.



• SOLDER POTS

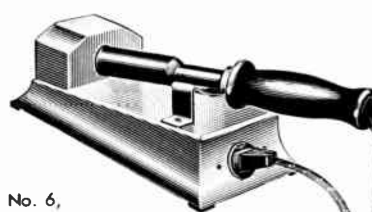
Ruggedly constructed cast iron pots, with easily replaceable elements.

Model No. 12—¾ lb. cap. 200 watt Net **$4.50**

Model No. 36—2¼ lbs. cap. 250 watt. Net **$5.50**

Model No. 60—3¾ lbs. cap. 325 watt. Net **$6.50**

• Temperature Control Stand



This control stand is thermostatically actuated by the tip temperature of the iron. This is the only logical way to control the temperature of a soldering iron, for there is too much lag between element and tip temperature for the application of a thermostat to the element, whether in an iron or a control stand. Cat. No. 5, irons up to 1″ dia. tip; Cat. No. 6, irons 1″ to 1⅝″ dia. tips. List price **$6.50**

ELECTRIC SOLDERING IRON CO., Inc.

2550 WEST ELM STREET **DEEP RIVER, CONN., U. S. A.**

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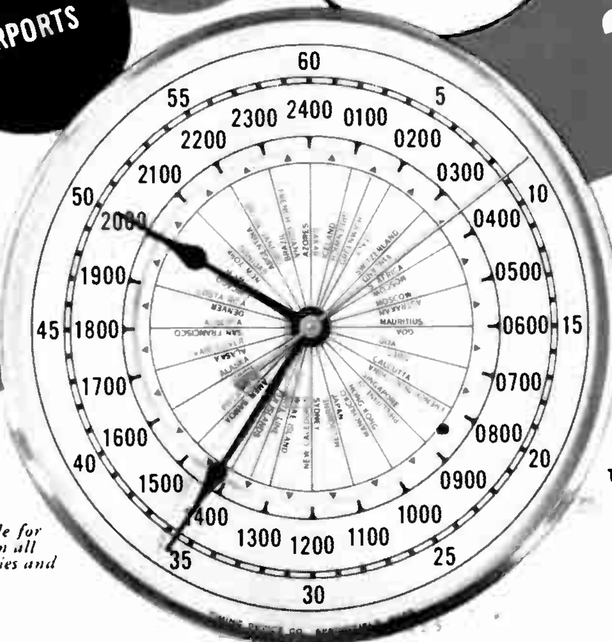
Here's the 2400-HOUR CLOC[K] with world-wide tim[e]



Available for Export in all frequencies and voltages

Here's a clock which not only gives accurate, easy to read 2400-hour time, in your own time zone, but also — clearly and instantaneously — the correct time in every time zone all over the world.

TD-2400

Specifications

MOTOR:
Synchronous clock motor (self-starting), AC operated.*

CASE:
10″ diameter — gray wrinkle finish, chrome-plated bezel.

HANDS:
Black, with red sweep second-hand.

DIALS:
10″ diameter — shows minutes, seconds and 2400 hours.

6″ diameter — rotating inner dial, frictionally attached to hour hand, in red and blue; shows time directly in all time zones.

CRYSTAL:
Convex annealed glass.

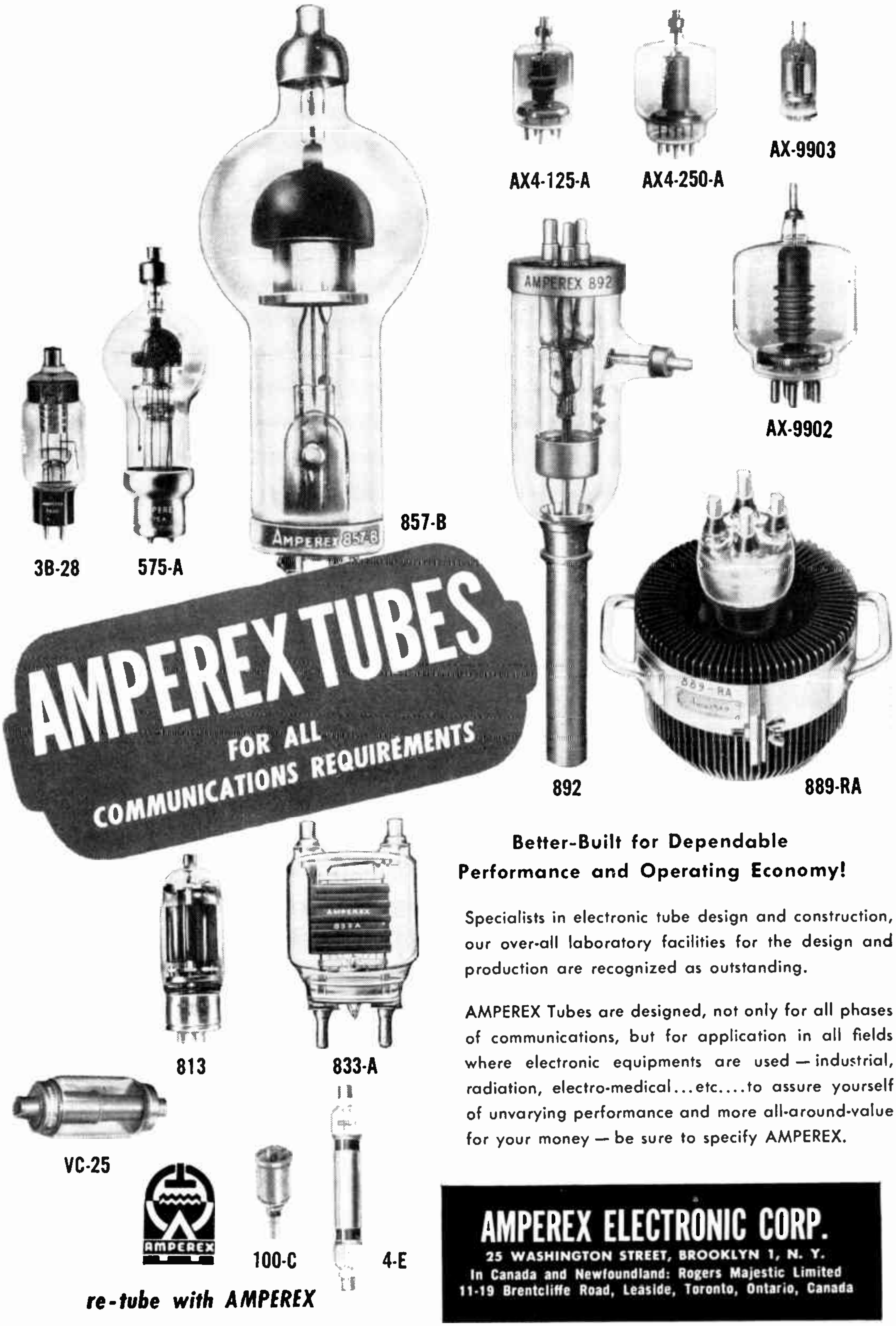
**standard 120-volt, 60-cycle; other voltages and frequencies on special orders.*

Check this list of features:

- Electric, self-starting.
- World-wide time — shows time directly and clearly in every time zone.
- Inner dial frictionally attached to hour hand — easily adjusted for use in any time zone.
- Big, 10-inch dial, lithographed in color; key cities and localities clearly shown.
- Sweep second-hand.
- Wall or panel mounting.

TIMING DEVICES CO.
SPRINGFIELD, MASS.

Manufacturers of clocks and special timing device[s] of all kinds. Sold only through established jobber[s]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If it's a problem calling for PRECISION POTENTIOMETERS

Bring it to Helipot

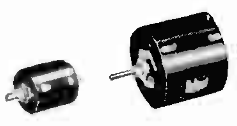
For many years The HELIPOT Corporation has been a leader in the development of advanced types of potentiometers. It pioneered the *helical* potentiometer—the potentiometer now so widely used in computer circuits, radar equipment, aviation devices and other military and industrial applications. It pioneered the DUODIAL*—the turns-indicating dial that greatly simplifies the control of multiple-turn potentiometers and other similar devices. And it has also pioneered in the development of many other unique potentiometric advancements where highest skill coupled with ability to mass-produce to close tolerances have been imperative.

In order to meet rigid government specifications on these developments—and at the same time produce them economically—HELIPOT* has perfected unique manufacturing facilities, including high speed machines capable of winding extreme lengths of resistance elements employing wire even less than .001" diameter. These winding machines are further supplemented by special testing facilities and potentiometer "know-how" unsurpassed in the industry.

So if you have a problem requiring *precision potentiometers* your best bet is to bring it to The HELIPOT Corporation. A call or letter outlining your problem will receive immediate attention!

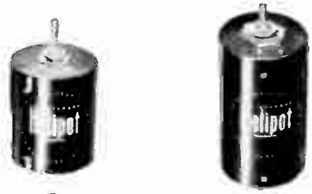
*Trade Marks Registered

In this panel are illustrated standard models of HELIPOT multi-turn and single-turn precision potentiometers—available in a wide range of resistances and accuracies to fulfill the needs of nearly any potentiometer application. The Beckman DUODIAL is furnished in two designs and four turns-ratios, to add to the usefulness of the HELIPOT by permitting easy and rapid reading or adjustment.



A B C

MODELS A, B, & C HELIPOTS
A—10 turns, 46" coil, 1-13/16" dia., 5 watts—resistances from 10 to 300,000 ohms.
B—15 turns, 140" coil, 3-5/16" dia., 10 watts—resistances from 50 to 500,000 ohms.
C—3 turns, 13-1/2" coil, 1-13/16" dia., 3 watts—resistances from 5 to 50,000 ohms.
—Ask for Bulletin 104—



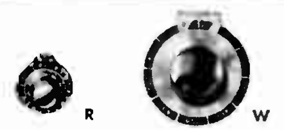
D E

MODELS D AND E HELIPOTS
Provide extreme accuracy of control and adjustment, with 9,000 and 14,400 degrees of shaft rotation.
D—25 turns, 234" coil, 3-5/16" dia., 15 watts—resistances from 100 to 750,000 ohms.
E—40 turns, 373" coil, 3-5/16" dia., 20 watts—resistances from 200 ohms to one megohm.
—Ask for Bulletin 104—

F G

MODELS F AND G PRECISION SINGLE-TURN POTENTIOMETERS
Feature both continuous and limited mechanical rotation, with maximum effective electrical rotation. Versatility of designs permit a wide variety of special features.
F—3-5/16" dia., 5 watts, electrical rotation 359°—resistances 10 to 100,000 ohms.
G—1-5/16" dia., 2 watts, electrical rotation 356°—resistances 5 to 20,000 ohms.
—Ask for Bulletin 105—

LABORATORY MODEL HELIPOT
The ideal resistance unit for use in laboratory and experimental applications. Also helpful in calibrating and checking test equipment. Combines high accuracy and wide range of 10-turn HELIPOT with precision adjustability of DUODIAL. Available in eight stock resistance values from 100 to 100,000 ohms, and other values on special order.
—Ask for Bulletin 106—



R W

MODELS R AND W DUODIALS
Each model available in standard turns-ratios of 10, 15, 25 and 40 to 1. Inner scale indicates angular position of HELIPOT sliding contact, and outer scale the helical turn on which it is located. Can be driven from knob or shaft end.
R—2" diameter, exclusive of index.
W—4-3/4" diameter, exclusive of index. Features finger hole in knob to speed rotation.
—Ask for Bulletins 104 and 114—

The versatility of the potentiometer designs illustrated above permit a wide variety of modifications and features, including double shaft extensions, ganged assemblies, the addition of a multiplicity of taps, variation of both electrical and mechanical rotation, special shafts and mounting bushings, high and low temperature operation, and close tolerances on both resistance and linearity. Examples of potentiometers modified for unusual applications are pictured at right.

A C

3-GANGED MODEL A HELIPOT AND DOUBLE SHAFT MODEL C HELIPOT
All HELIPOTS, and the Model F Potentiometer, can be furnished with shaft extensions and mounting bushings at each end to facilitate coupling to other equipment.
The Model F, and the A, B, and C HELIPOTS are available in multiple assemblies, ganged at the factory on common shafts, for the control of associated circuits.

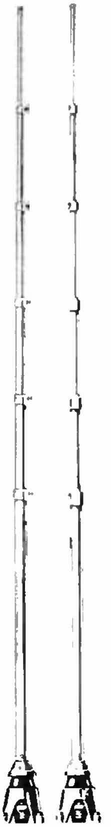
B F

MULTITAPPED MODEL B HELIPOT AND 4-GANGED TAPPED MODEL F
This Model B HELIPOT contains 28 taps, placed as required at specified points on coil. The Four-Gang Model F Potentiometer contains 10 taps on each section. Such taps permit use of padding resistors to create desired nonlinear potentiometer functions, with advantage of flexibility, in that curves can be altered as required.

THE Helipot CORPORATION, SOUTH PASADENA 10, CALIFORNIA

PREMAX

VERTICAL AND MOBILE ANTENNAS
ELEMENTS--MOUNTINGS--ACCESSORIES



Steel Alumi'm
Less Base

Premax Tubular Vertical Antennas are fully collapsing and adjustable, yet give exceptionally efficient, dependable performance under most severe conditions. Will withstand ordinary stresses but should be supported by guys or standoff insulators against abnormal winds. In 6′ to 35′ heights, in aluminum and steel.

Weather-Resistant Steel Antennas

No.	Description	Length Extended	Length Collapsed	O.D. Base	I.D. Base	Weight each, lbs.
112-M	2-sec. telescop'g	11′8″	6′1″	.656″	.556″	4
318-M	3-sec. telescop'g	17′3″	6′2″	.875″	.775″	7
224-M	4-sec. telescop'g	22′9″	6′3″	1.063″	.963″	11
130-M	5-sec. telescop'g	28′3″	6′1″	1.250″	1.150″	15
136-M	6-sec. telescop'g	33′9″	6′5″	1.500″	1.400″	20

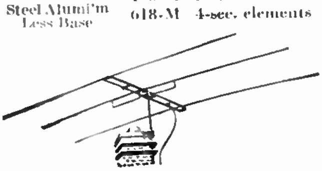
Aluminum Antennas

No.	Description	Length Extended	Length Collapsed	O.D. Base	I.D. Base	Weight each, lbs.
AL-106	1-piece taper rod	6′3″	6′3″	.313″	....	¼
AL-312	2-sec. telescop'g	12′4″	6′4″	.500″	.334″	1½
AL-518	3-sec. telescop'g	18′5″	6′1″	.750″	.584″	3
AL-324	4-sec. telescop'g	24′4″	6′4″	1.000″	.834″	5
AL-530	5-sec. telescop'g	30′0″	6′5″	1.250″	1.084″	7
AL-535	6-sec. telescop'g	35′8″	6′5″	1.500″	1.310″	12

CORULITE ELEMENTS for Beam Arrays

Premax Corulite Elements meet the need for sturdy, light-weight elements in horizontal arrays and similar applications. Exceptionally light weight, yet they provide the needed strength and rigidity so essential in horizontal installations — and at an extremely low cost. The special steel tubing is a Premax development to insure unusual stiffness and strength. Heavily electroplated to insure corrosion resistance and high electrical conductivity. Fully adjustable to any desired length.

No.	Description	Length Extended	Length Collapsed	O.D. Base	Recommended For	Weight per pair
108-M	2-sec. elements	8′2″	4′7″	.750″	10-meter	2 lbs.
618-M	4-sec. elements	17′0″	5′3″	1.000″	20-meter	5½ lbs.



ROTARY BEAM KIT

Complete Rotary Beam Kit No. RB-6309 for 6, 10 and 11 meters, includes aluminum frame, 3 pairs Elements, T-Match Accessories and necessary hardware. Weighs only 20 lbs.

BASE-LOADED MOBILE

Mobile "75" Base Loaded Antenna vastly improves radiation characteristics over standard "whip" types. Has 6 db. gain, equal to quadrupling transmitting power. Greatly improves effectiveness and range, both on transmission and reception. Has extra-long, space-wound, base-loaded inductor, topped by vertical whip. A low-cost solution to mobile 75-meter troubles. Base Extension also available to convert to center-loaded giving additional 2 to 3 db. gain.

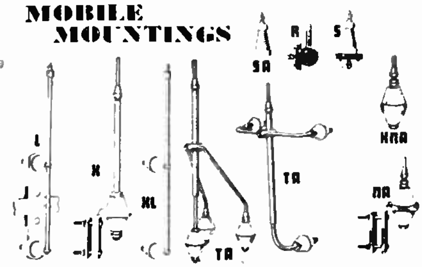
WHIP TYPE ANTENNAS

Three outstanding 1-piece tapered "whip" styles designed for maximum strength and required flexibility to meet most exacting requirements. Also two splendid low-cost oil-tempered, special-formula solid steel "whip" Antennas. Available in cadmium-plated steel, stainless steel, aluminum and high-tempered steel in 72″, 84″ and 96″ lengths.

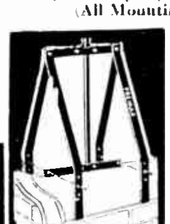
ROTOMOUNT

Of heavy sheet steel, spot-welded, Rotates full 360°. Platform size 10″ x 12″, supported by 7″ ball thrust bearing; ⅝″ center shaft leadin opening. Wt. 17 lbs.

MOBILE MOUNTINGS



No. TA Mounting for trunk, panel or roof.
No. K Bumper Mounting permits 10″ adjustment.
No. XL Mounting for panel of car.
No. L Bumper Mounting, 10″ adjustment.
No. NNA Fender Mounting for flat surfaces.
No. NA Bumper Mounting, 2″ adjustment.
No. R Universal Mounting; split-ball type; adjustable for any position on car.
No. S Roof Mounting; spring style.
No. SA Spring Adaptor for use with any mount.
(All Mountings fit any ¼-inch whip antenna)



Parapet Support

For masts or radiator on parapets or fire walls up to 20″. Fits mast ⅞″ to 2″ diameter. An ideal support. No. PS-18 Parapet Support, weight 8 lbs.

PREMAX INSULATORS AND MOUNTINGS

Type 1

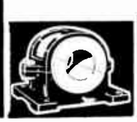
Type 2

Type 6

Bushing

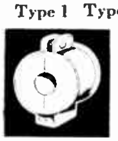
7

13-S



Type 8-C

Type 9-C

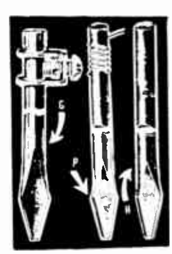


Type 10-C

10-S

No. 1 Base Insulator, galv. iron or bronze; fits ¾″ to 1⅜″ I.D. masts.
No. 2 Base Insulator, galv. iron or bronze; fits ¾″ or 13/16″ I.D. masts.
No. 6 Base Insulator, galv. or bronze for rooftop, deck or marine; fits ¾″ to 1⅜″ I.D.
Bushing, brown glaze porcelain with galv. flange for roof or deck; fits ¾″, 1¼″, 1¾″ O.D.
No. 7 Standoff Insulator for supporting verticals or as element mount; fits ⅝″ to 1″.
No. 8 Insulated Mounting Clamp for horizontals or verticals; fits ⅝″ to 1″ O.D.
No. 9-C Insulated Mounting Clamp for horizontals or verticals; fits ⅝″ to 1″ O.D.
No. 10-C Insulated Mounting Clamp for horizontals or verticals; fits ⅝″ to 1″ O.D
No. 10-S Heavy Duty Standoff; chrome plated; fits ⅞″ to 1¼″ O.D.
No. 13-S Heavy Duty Standoff; plain or hinged cap; fits ¾″ to 1½″ O.D.

GROUND RODS



Made of copper-plated steel, in ⅜″, ½″ and ⅝″ diameters and in 4′, 5′, 6′ and 8′ lengths. One end pointed for easy driving. (g) Screw clamp, (P) Pigtail, (H) Drilled hole types available.

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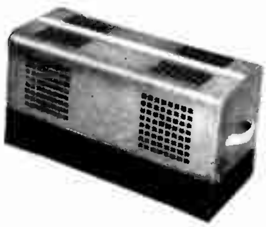

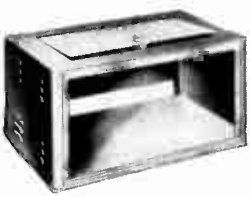

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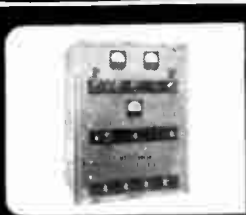

New WRL "400" GLOBE KING

A versatile, advanced design transmitter that gives efficient performance on all bands—10 to 160 on phone and CW. 350 watt phone — 400 watt CW. Provisions for ECO. Complete with one set of coils.

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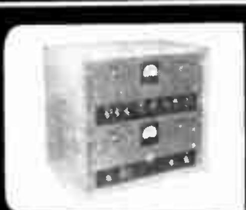

WRL 175 Watt GLOBE CHAMPION

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Capable of 25 watt input on phone and 40 watt input on CW on all bands from 1500 KC thru 30 megacycles. Band switching for any 3 bands. A proven rig—thousands in use throughout the world.

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less tubes, coils, and meter
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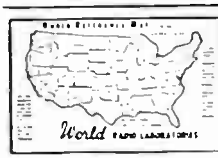

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In keeping with its policy of providing all services within its power, The American Radio Relay League makes available to amateurs and would-be amateurs literature properly prepared to present in the best form all available information pertaining to amateur radio. The fact that its offices are the national and international headquarters of radio amateurs, makes League publications authoritative, complete, up-to-the-minute; written from a thoroughly practical amateur's point-of-view. These publications are frequently revised and augmented to keep abreast of the fast-changing field. All are printed in the familiar *QST* format which permits thorough but economical presentation of the information. Most of the publications and supplies described in the following pages are handled by your dealer for your convenience.

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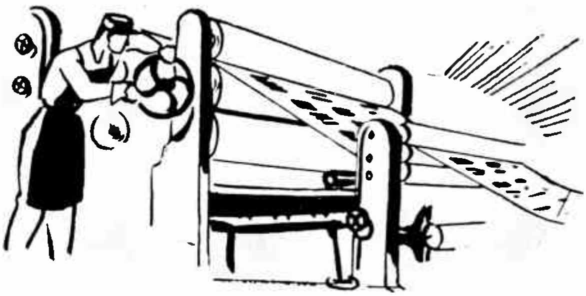
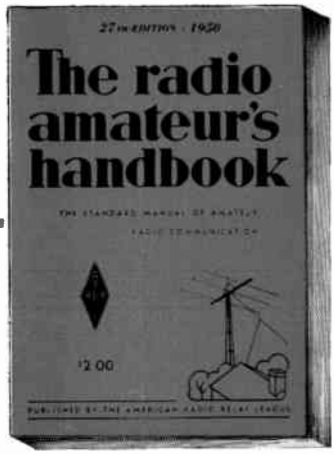


THE OFFICIAL MAGAZINE OF THE AMERICAN RADIO RELAY LEAGUE

QST faithfully and adequately reports each month the rapid development which makes Amateur Radio so intriguing. Edited in the sole interests of the members of The American Radio Relay League, who are its owners, *QST* treats of equipment and practices and construction and design, and the romance which is part of Amateur Radio, in a direct and analytical style which has made *QST* famous all over the world. It is essential to the well-being of any radio amateur. *QST* goes to every member of The American Radio Relay League and membership costs $4.00 in the United States and Possessions, $4.50 in the Dominion of Canada, $5.00 in all other countries. Elsewhere in this book will be found an application blank for A.R.R.L. membership.

For thirty-five years (and thereby the oldest American radio magazine) *QST* has been the "bible" of Amateur Radio.



The *Handbook* tells the things which are needed for a comprehensive understanding of Amateur Radio. From the story of how Amateur Radio started through an outline of its wide scope of the present—from suggestions on how to learn the code through explanations of traffic-handling procedure and good operating practices—from electrical and radio fundamentals through the design, construction, and operation of amateur equipment—this book covers the subject thoroughly. It includes the latest and the best information on everything in Amateur Radio.

$2.00 U.S.A. proper
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$3.00 U.S.A. proper
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The standard elementary guide for the prospective amateur. Features equipment which, although simple in construction, conforms in every detail to present practices. The apparatus is of a thoroughly practical type capable of giving long and satisfactory service—while at the same time it can be built at a minimum of expense. The design is such that a high degree of flexibility is secured, making the various units fit into the more elaborate station layouts which inevitably result as the amateur progresses. Complete operating instructions and references to sources of detailed information on licensing procedure are given.

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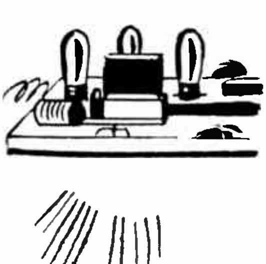

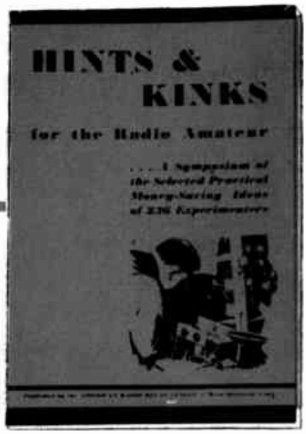

Ama-
teurs are
noted for their
ingenuity in overcom-
ing by clever means the minor
and major obstacles they meet
in their pursuit of their chosen hobby.
An amateur must be resourceful and a
good tinkerer. He must be able to make a
small amount of money do a great deal for him.
He must frequently be able to utilize the contents of
the junk box. Hints and Kinks is a compilation of
hundreds of good ideas which amateurs have
found helpful, including conversion pro-
cedures for war surplus. It will return its
cost many times in money savings—
and it will save hours of time.

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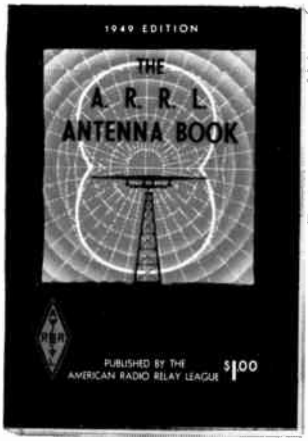
A
compre-
hensive manual
of antenna design and
construction, by the head-
quarters staff of the American
Radio Relay League. Sixteen chap-
ters, profusely illustrated. Both the theory
and the practice of all types of antennas
used by the amateur, from simple doublets
to multi-element rotaries, including long wires,
rhomboids, vees, phased systems, v.h.f systems,
etc. Feed systems and their adjustment.
Construction of masts, lines and rotating
mechanisms. The most comprehensive
and reliable information ever
published on the subject.



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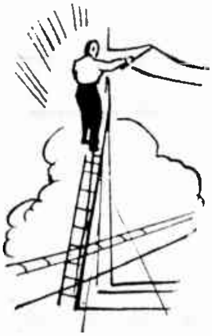
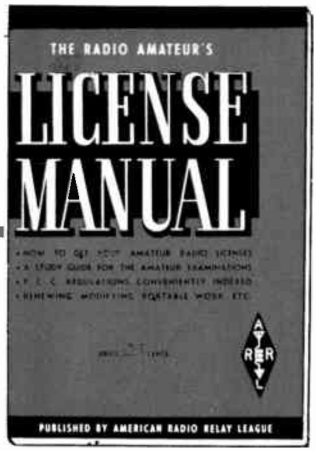


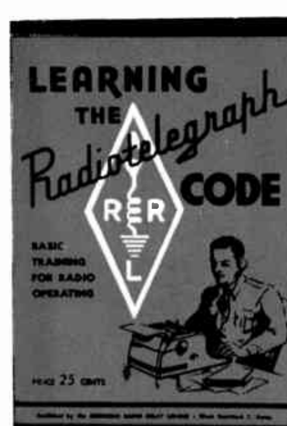


To obtain an amateur operator's license you must pass a government examination. The License Manual tells how to do that—tells what you must do and how to do it. It makes a simple and comparatively easy task of what otherwise might seem difficult. In addition to a large amount of general information, it contains questions and answers such as are asked in the government examinations. If you know the answers to the questions in this book, you can pass the examination without trouble.

25c

Postpaid

This booklet is designed to train students to handle code skillfully and with precision, both in sending and in receiving. Employing a novel system of code-learning based on the accepted method of *sound* conception, it is particularly excellent for the student who does not have the continuous help of an experienced operator or access to a code machine. It is similarly helpful home-study material for members of code classes. Adequate practice material is included for classwork as well as for home-study. There are also helpful data on high-speed operation, typewriter copy, general operating information—and an entire chapter on tone sources for code practice, including the description of a complete code instruction table with practice oscillator.



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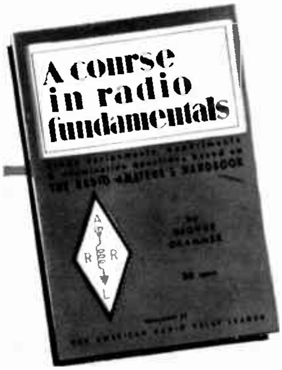


"A Course In Radio Fundamentals" is a study guide, examination book and laboratory manual. Its text is based on the "Radio Amateur's Handbook." As a text, this book greatly smooths the way for the student of the technicalities of radio. It contains interesting study assignments, experiments and examination questions for either class or individual instruction. It describes in detail 40 experiments with simple apparatus giving a complete practical knowledge of radio theory and design.

50c

Postpaid Anywhere

The current edition of the A.R.R.L. Amateur Radio MAP, entirely different in conception and design from any other map, contains a wealth of information useful to the radio amateur. A special type of projection made by Rand, McNally to A.R.R.L. specifications, it gives great circle distance measurements in miles or kilometers within an accuracy of 2%. The map shows principal cities of the world; local time zones; WAC divisions; more than 265 countries, indexed; and amateur prefixes throughout the world. Large enough to be easily readable from your operating position, the map is printed in six colors on heavy paper, 30 x 40 inches.

$2.00

Postpaid Anywhere

STATION OPERATING *Sup*

LOG BOOK

As can be seen in the illustration, the log page provides spaces for all facts pertaining to transmission and reception, and is equally as useful for portable or mobile operation as it is for fixed. The log pages with an equal number of blank pages for notes and a sheet of graph paper are spiral bound, permitting the book to be folded back flat at any page, requiring only the page size of 8½ x 11 on the operating table. In addition, a number sheet, with A.R.R.L. Numbered Texts printed on back, for traffic handlers, is included with each book.

50c per book

Official Radiogram Forms

The radiogram blank is designed to comply with the proper order of transmission. All blocks for fill-in are properly spaced for use in typewriter. It has a heading that you will like. Radiogram blanks, 8½ x 7¼, lithographed in green ink, and padded 100 blanks to the pad, 35c per pad, postpaid.

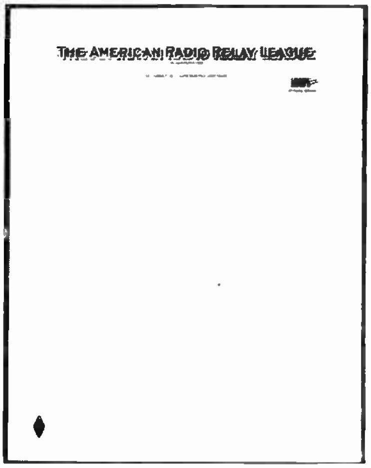
Message Delivery Cards

Radiogram delivery cards embody the same design as the radiogram blank and are available in two styles—on stamped government postcard, 2c each; unstamped, 1c each.

The operating supplies shown on this page have been designed by the A.R.R.L. Communications Department.

lies

MEMBERSHIP



Members' Stationery

Members' stationery is lithographed on standard 8½ x 11 bond paper which every member should be proud to use for his radio correspondence.

100 Sheets, **$1.00** 250 Sheets, **$1.50**

500 Sheets
$2.50
postpaid

In the January, 1920 issue of ***QST*** there appeared an editorial requesting suggestions for the design of an A.R.R.L. emblem—a device whereby every amateur could know his brother amateur when they met. In the July, 1920 issue the design was announced—the familiar diamond that greets you everywhere in Ham Radio. For years it has been the unchallenged emblem of amateur radio.

The League Emblem, with gold border and lettering, and with black enamel background, is available in either pin (with safety clasp) or screw-back type In addition, there are special colors for Communications Department appointees. • Red enameled background for the SCM. • Blue enameled background for the ORS or OPS.

50c each *postpaid*

The Emblem Cut. A mounted printing electrotype, ⅝" high, for use by members on amateur printed matter, letterheads, cards, etc.

$1.00
each
postpaid

Lightning Calculators

Aware of the practical bent of the average amateur and knowing of his limited time, the League, under license of the designer, W. P. Koechel, has made available these calculators to obviate the tedious and sometimes difficult mathematical work involved in the design and construction of radio equipment. The lightning calculators are ingenious devices for rapid, certain and simple solution of the various mathematical problems which arise in radio and allied work. They make it possible to read direct answers without struggling with formulas and computations. They are tremendous time-savers for amateurs, engineers, servicemen and experimenters. Their accuracy is more than adequate for the solution of practical problems, and is well within the limits of measurement by ordinary means. Each calculator has on its reverse side detailed instructions for its use; the greatest mathematical ability required is that of dividing or multiplying simple numbers. They are printed in several colors. You will find lightning calculators the most useful gadgets you ever owned.

TYPE A

Radio Calculator

This calculator is useful for the problems that confront the amateur every time he builds a new rig or rebuilds an old one or winds a coil or designs a circuit. It has two scales for physical dimensions of coils from one-half inch to five and one-half inches in diameter and from one-quarter to ten inches in length; a frequency scale from 400 kilocycles through 150 megacycles; a wavelength scale from two to 600 meters; a capacity scale from 3 to 1,000 micro-microfarads; two inductance scales with a range of from one microhenry through 1,500; a turns-per-inch scale to cover enameled or single silk covered wire from 12 to 35 gauge, double silk or cotton covered from 0 to 36 and double cotton covered from 2 to 36. Using these scales in the simple manner outlined in the instructions on the back of the calculator, it is possible to solve problems involving frequency in kilocycles, wavelength in meters, inductance in microhenrys and capacity in microfarads, for practically all problems that the amateur will have in designing—from high-powered transmitters down to simple receivers. Gives the direct reading answers for these problems with accuracy well within the tolerances of practical construction. **$1.00** *Postpaid.*

TYPE B

Ohm's Law Calculator

This calculator has four scales: a power scale from 10 microwatts through 10 kilowatts, a resistance scale from .01 ohm through 100 megohms, a current scale from 1 microampere through 100 amperes, a voltage scale from 10 microvolts through 10 kilovolts.

With this concentrated collection of scales, calculations may be made involving voltage, current, and resistance, and can be made with a single setting of a dial. The power or voltage or current or resistance in any circuit can be found easily if any two are known. This is a newly-designed Type B Calculator which is more accurate and simpler to use than the justly-famous original model. It will be found useful for many calculations which must be made frequently but which are often confusing if done by ordinary methods. All answers will be accurate within the tolerances of commercial equipment. **$1.00** *Postpaid.*