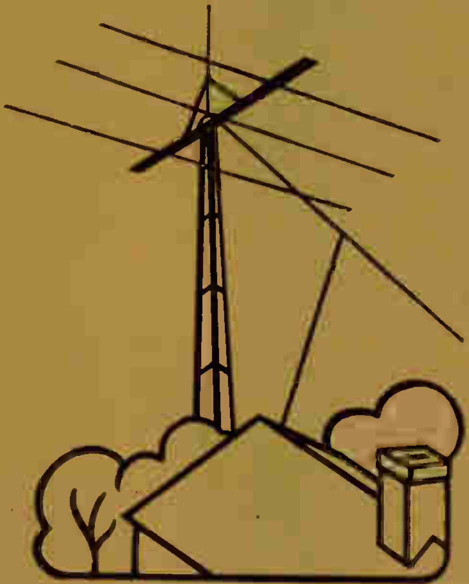
28TH EDITION · 1951

The radio amateur's handbook

THE STANDARD MANUAL OF AMATEUR
RADIO COMMUNICATION

$2.50
U.S.A. Proper

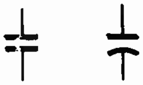


PUBLISHED BY THE AMERICAN RADIO RELAY LEAGUE

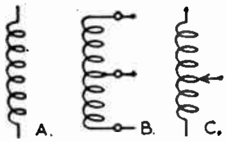
SCHEMATIC SYMBOLS USED IN CIRCUIT DIAGRAMS

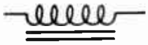

ANTENNA

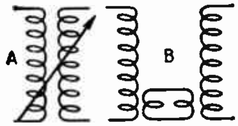

GROUND

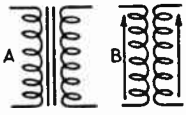

FIXED CONDENSER
(See Footnote 1)


VARIABLE OR ADJUSTABLE CONDENSER
A- Single-section
B- Split-stator
(See footnote 2)


AIR-CORE INDUCTOR
A- Fixed coil or r.f. choke
B- Coil with fixed tap
C- Coil with variable tap
(Small circles indicate plug-and-jack or binding-post terminals)


IRON-CORE INDUCTOR
OR CHOKE


A-AIR-CORE TRANSFORMER OR INDUCTIVELY-COUPLED COILS
(Arrow used only if coupling is variable)
B- LINK-COUPLED COILS


IRON-CORE TRANSFORMERS
A- Laminated core
B- Powdered-iron core
(Arrows indicate variable core or permeability tuning)


FIXED RESISTOR


VARIABLE RESISTOR.
POTENTIOMETER, VOLTAGE
DIVIDER, RHEOSTAT, ETC.


WIRING-DIAGRAM DEVICES
A- Wires connected
B- Wires not connected


TWISTED-PAIR CABLE

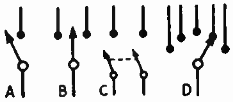

COAXIAL CABLE


SHIELDED WIRE OR CABLE


SHIELDING


FARADAY SHIELD

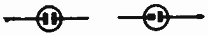

TERMINALS
with appropriate labels

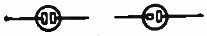

SWITCHES
A- S.P.S.T. C- D.P.S.T.
B- S.P.D.T. D-Rotary Multipoint


KEY

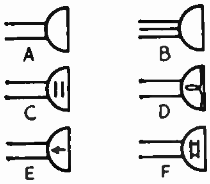

JACK

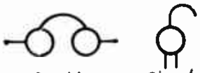

PLUG

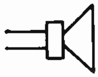

POWER PLUGS
A- Non polarized
B- Polarized


Nonpolarized and polarized power receptacles

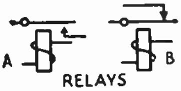

FUSE

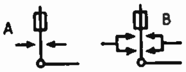

MICROPHONES
A- Single-button D- Dynamic
B- Double-button E- Velocity
C- Condenser F- Crystal


Double *Single*
HEADPHONES


LOUDSPEAKER


BUZZER

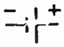

RELAYS
A- Normally-open
B- Normally-closed


VIBRATORS
A- Non rectifying
B- Self-rectifying


METER
(with # = proper identification - V, MA, etc.)

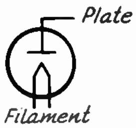

MOTOR


BATTERY

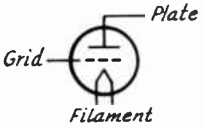

SINGLE CELL


RECTIFIER
(Usually dry-disk)

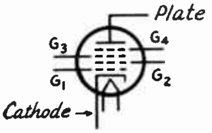
•*(dot)*
INDICATES GASEOUS TUBE



DIODE VACUUM TUBE



TRIODE VACUUM TUBE



MULTIGRID VACUUM TUBE
The grids are usually numbered, G₁ being that closest to the cathode


FILAMENT
OR HEATER


CATHODE


PHOTOELECTRIC
CATHODE


COLD
CATHODE


GRID
(Also beam-confining or beam-forming electrode)


PLATE


DIODE
PLATE


ANODES


ELECTRON-RAY
TUBE TARGET
ANODES


CATHODE-RAY
TUBE DEFLECTING
PLATES


LAMPS
A- Panel or dial
B- Illuminating


NEON BULB or VOLTAGE REGULATOR
(VR) TUBE


CRYSTALS
A- Piezoelectric
B- Detector

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[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, CONN., U.S.A.

1951

Twenty-Eighth Edition

Twenty-Eighth Edition

THE RUMFORD PRESS
Concord, New Hampshire, U. S. A.

Foreword

This twenty-eighth edition of *The Radio Amateur's Handbook* is the latest of a series extending over twenty-five 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 changes that have taken place in the technical practices of amateur radio during the past year are, as always, reflected in the present edition. A considerable amount of new equipment in all categories — transmitting, receiving, v.h.f., measurements — appears throughout the book. Continuing the trend of the past few years, all transmitting equipment has been designed with the reduction of harmonics in the television broadcasting bands as a primary feature, and in view of the large number of television transmitting stations now in operation, the problems of amateur interference with this service are given special attention. As compared with previous editions, the plan of the book has been modified somewhat: Radiotelephony, formerly covered in its entirety in one chapter, is now broken down into several chapters dealing with audio equipment, various basic types of modulation, and single-sideband; similarly, transmission lines and antennas are now treated in separate chapters. It is the thought that by this means information on a specialized phase of a general subject will be made more readily accessible.

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, 1950

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 shortwave radio. Scattered over the globe are more than 125,000 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 90,000, 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 K6BJ, 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 eight 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 cooperative 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 cooperation 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 World War II, 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.

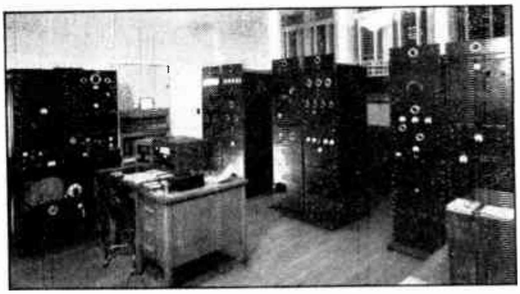
Emergency relief, expedition contact, experimental work and countless instances of other forms of public service — rendered, as they always have been and always will be, without hope or expectation of material reward — made amateur radio an integral part of our peacetime national life. The importance 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 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. It represents the amateur in legislative matters.

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 144 Mc. must be adequately-filtered direct current. Emissions must be free from spurious radiations. The licensee must provide for measurement of the transmitter frequency and establish a procedure for checking it regularly. A complete log of station operation must be maintained, with specified data. The station license also authorizes the holder to operate portable and mobile stations on certain frequencies, subject to further regulations. An amateur station may be operated only by an amateur operator licensee, but any licensed amateur operator may operate any amateur station. All radio licensees are subject to penalties for violation of regulations.

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

● LEARNING THE CODE

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

A	didah	**N**	dahdit
B	dahdididit	**O**	dahdahdah
C	dahdidahdit	**P**	didahdahdit
D	dahdidit	**Q**	dahdahdidah
E	dit	**R**	didahdit
F	dididahdit	**S**	dididit
G	dahdahdit	**T**	dah
H	didididit	**U**	dididah
I	didit	**V**	didididah
J	didahdahdah	**W**	didahdah
K	dahdidah	**X**	dahdididah
L	didahdidit	**Y**	dahdidahdah
M	dahdah	**Z**	dahdahdidit
1	didahdahdahdah	6	dahdidididit
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).

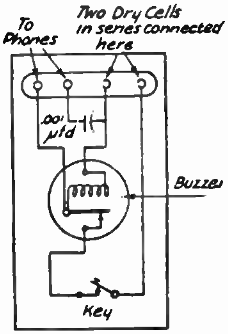


Fig. 1-2 — The headphones are connected across the coils of the buzzer, with a condenser in series. If the value shown gives an excessively loud signal, it may be reduced to 470 $\mu\mu$fd. or 220 $\mu\mu$fd.

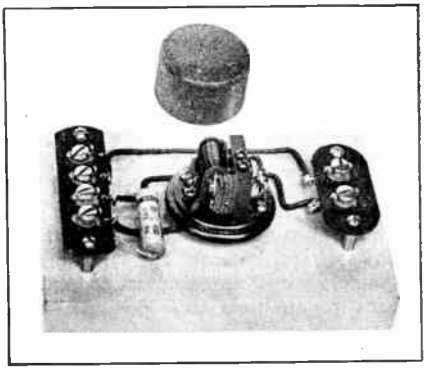


Fig. 1-3 — The cover of the buzzer unit has been removed in this view of the buzzer code-practice set.

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

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

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.

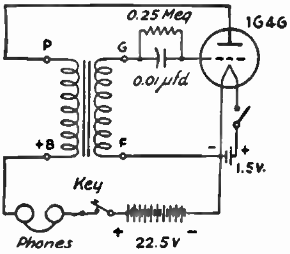


Fig. 1-4 — Wiring diagram of a simple vacuum-tube audio-frequency oscillator for use as a code-practice set.

● 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.

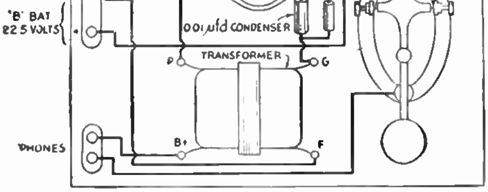


Fig. 1-5 — Layout of the audio-oscillator code-practice set. All parts may be mounted on a wooden baseboard, approximately 5 by 7 inches in size.

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

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

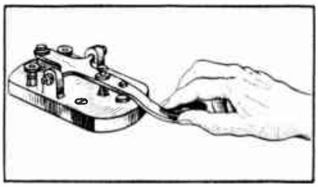


Fig. 1-6 — This sketch illustrates the correct position of the hand and fingers for good sending with a telegraph key.

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

● THE AMATEUR BANDS

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

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. A0 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; either c.w. or voice may be used.

Area	*Band, kc*	*Power (watts)* *Day*	 *Night*
Mississippi River to East Coast U.S. (except Florida and states bordering Gulf of Mexico)	1800–1825 kc 1875–1900 kc	500	200
Mississippi River to West Coast U.S. (except states bordering Gulf of Mexico)	1900–1925 kc. 1975–2000 kc	*500	*200
Florida and states bordering Gulf of Mexico	1800–1825 kc 1875–1900 kc	200	No operation
Puerto Rico and Virgin Islands	1900–1925 kc 1975–2000 kc	500	50
Hawaiian Islands	1900–1925 kc 1975–2000 kc	500	200

* Except in State of Washington where daytime power limited to 200 watts and night time power to 50 watts.

The 1947 International Radio Conference resulted in certain planned changes in present bands which may become effective some time in 1951. 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 1950 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.

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	— A0, 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	— A0, A1, A2, A3, A4, FM
220 -225	— A0, A1, A2, A3, A4, FM
420* -450*	— A0, A1, A2, A3, A4, A5, FM
1,215 - 1,300	— A0, A1, A2, A3, A4, A5, FM
2,300 - 2,450	A0, A1, A2, A3, A4, A5, FM, Pulse
3,300 - 3,500	
5,650 - 5,925	
10,000 -10,500	
21,000 -22,000	
All above 30,000	

* Peak antenna power must not exceed 50 watts.

● NEW LICENSE STRUCTURE

Also at press time announcement is imminently expected of a change in the license structure, the most pertinent aspect of which would be the creation of two new classes of license sometime during 1951. One would be a Novice Class, requiring the passing of an FCC examination including a code test at only 5 words per minute and a very simple written test in theory and regulation. The Novice Class license would have a term of only one year, and could not be renewed; its purpose would be to provide on-the-air training for newcomers wishing to become "full-fledged" amateurs.

As might be expected from the reduced requirements, the privileges available to the Novice Class licensee would be limited. Maximum input power probably would be 75 watts and crystal control would be required. Code could be used in 3700–3750 kc. and 26,960–27,230 kc., and either code or voice in 145–147 Mc.

For the newcomer to amateur radio who is particularly interested in the ultrahigh frequencies and micro-waves, it is expected there will also be a Technician Class license, with a 5 w.p.m. code test but the same written test as for present Class B license. Privileges would be all those in bands above 220 Mc. This would be a five-year term license, renewable in the usual manner.

Consult *QST* or write ARRL Headquarters in West Hartford, Conn., for any further data about these licenses.

CHAPTER 2

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 a practical reference section, covering those principles that are basic to all radio equipment. However, those who wish to use it as a course in elementary circuit theory will find that the material is arranged so it can be studied consecutively. Such study is, of course, valuable; but the chapter should prove even more valuable when frequently consulted for help in the solution of everyday problems.

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-

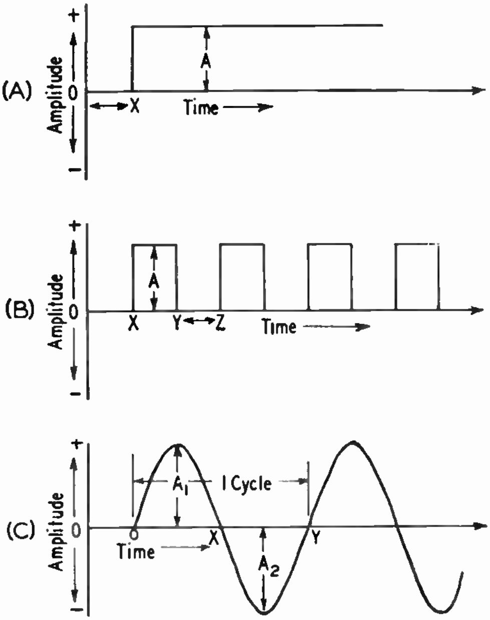


Fig. 2-1 — Three types of current flow. A — direct current; B — intermittent direct current; C — alternating current.

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.

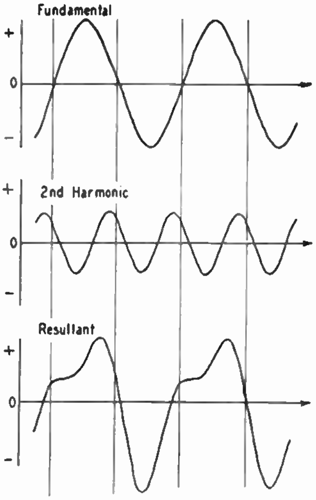


Fig. 2-2 — A complex waveform. A fundamental (top) and second harmonic (center) added together, point by point at each instant, result in the waveform shown at the bottom. When the two components have the same polarity at a selected instant, the resultant is the simple sum of the two. When they have 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 1000 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 the Miscellaneous Data chapter. 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 the Miscellaneous Data chapter 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 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.

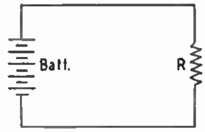


Fig. 2-3 — A simple circuit consisting of a battery and resistor.

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:

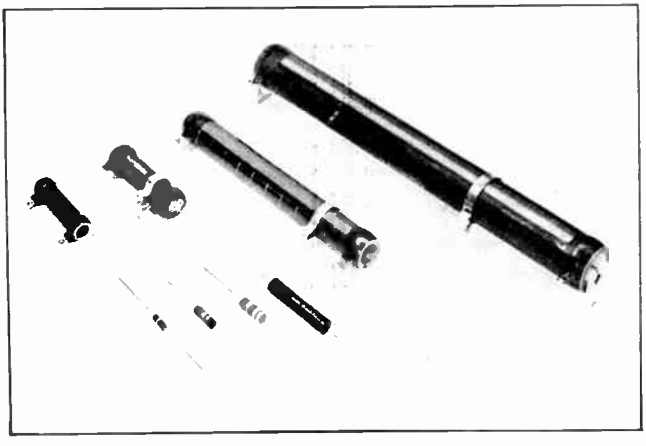
$$E = IR$$

(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. × 1000 = 50 milliamperes.

● SERIES AND PARALLEL RESISTANCES

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

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

TABLE 2-II
Conversion Values for Fractional and Multiple Units

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

Resistors in Series

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

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

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

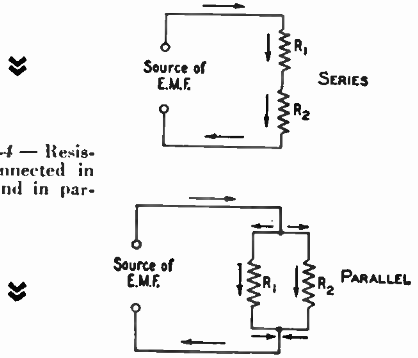


Fig. 2-4 — Resistors connected in series and in parallel.

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

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

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

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

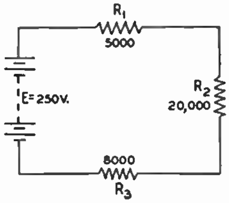


Fig. 2-5 — An example of resistors in series. The solution of the circuit is worked out in the text.

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

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

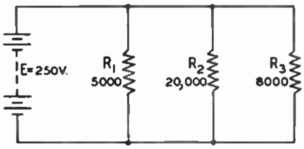


Fig. 2-6 — An example of resistors in parallel. The solution is worked out in the text.

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

$$R = \frac{R_1R_2}{R_1 + R_2}$$

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

$$R = \frac{R_1R_2}{R_1 + R_2} = \frac{500 \times 1200}{500 + 1200} = \frac{600{,}000}{1700}$$
$$= 353 \text{ ohms}$$

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

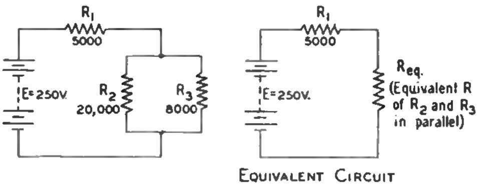
$$I = I_1 + I_2 + I_3 = 50 + 12.5 + 31.25$$
$$= 93.75 \text{ ma.}$$

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



EQUIVALENT CIRCUIT

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.)

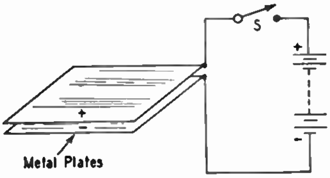


Fig. 2-8 — A simple condenser.

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).

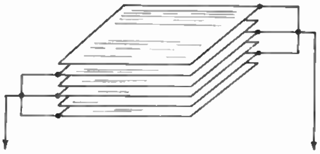


Fig. 2-9 — A multiple-plate condenser. Alternate plates are connected together.

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

The formula for calculating the capacitance of a condenser is:

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

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

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

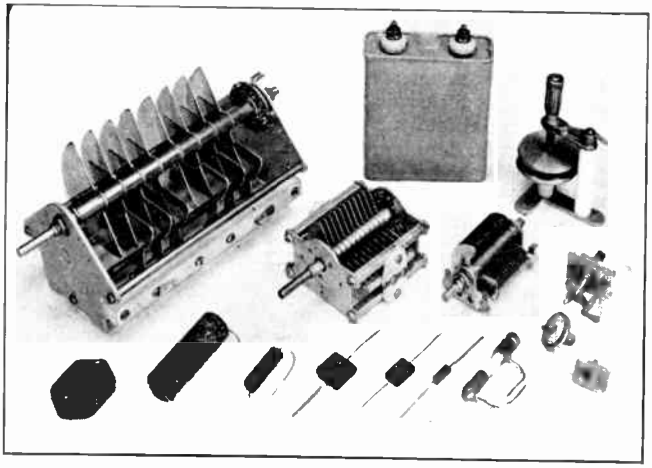
Example: A "variable" condenser has 7 semicircular plates on its rotor, the diameter of the semicircle being 2 inches. The stator has 6 rectangular plates, with a semicircular cut-out to clear the rotor shaft, but otherwise large enough to face the entire area of a rotor plate. The diameter of the cut-out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

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

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

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

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



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

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

Condensers in Radio

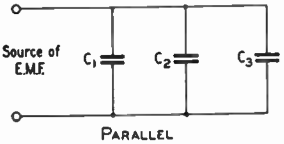
The types of condensers used in radio work differ considerably in physical size, construction, and capacitance. Some representative types are shown in the photograph. In "variable" condensers (almost always constructed with air for the dielectric) one set of plates is made movable with respect to the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

Voltage Breakdown

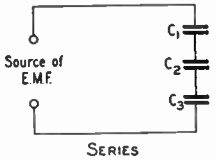
When a high voltage is applied to the plates of a condenser, a considerable force is exerted on the electrons and nuclei of the dielectric. Because the dielectric is an insulator the electrons do not become detached from atoms the way they do in conductors. However, if the force is great enough the dielectric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

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

● CONDENSERS IN SERIES AND PARALLEL



PARALLEL



SERIES

Fig. 2-10 — 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 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 μμfd.; 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.

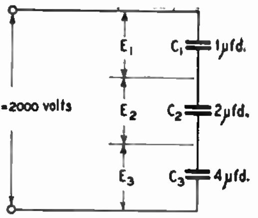


Fig. 2-11 — An example of condensers connected in series. The solution to this arrangement is worked out in the text.

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 Supplies) 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 (Miscellaneous Data chapter), 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:

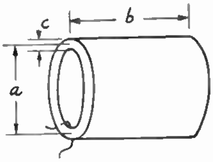


Fig. 2-12 — Coil dimensions used in the inductance formula.

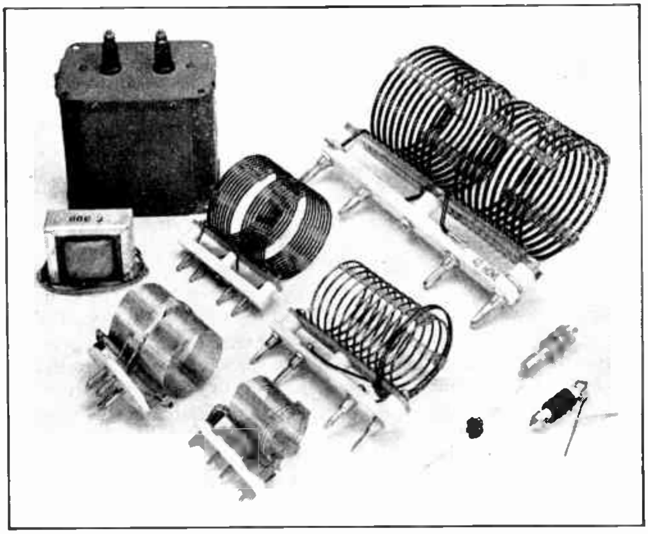
$$N = \sqrt{\frac{3a + 9b}{0.2a^2} \times L}$$

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,

$$N = \sqrt{\frac{(3 \times 1) + (9 \times 1.25)}{0.2 \times 1^2} \times 1}$$
$$= \sqrt{\frac{14.25}{0.2} \times 10} = \sqrt{712.5}$$
$$= 26.6 \text{ turns.}$$

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

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



◆

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

◆

● IRON-CORE COILS

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

Permeability

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

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

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

Eddy Currents and Hysteresis

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

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

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

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

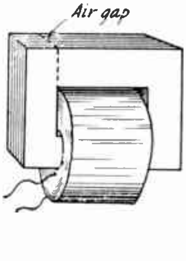


Fig. 2-13 — Typical construction of an iron-core coil. The small air gap prevents magnetic saturation of the iron and increases the inductance at high currents.

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

INDUCTANCES IN SERIES AND PARALLEL

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

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

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

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

and for two inductances in parallel,

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

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

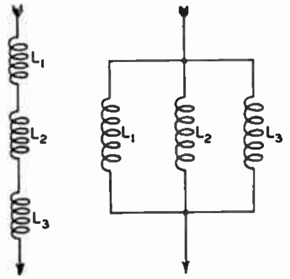


Fig. 2-14 — **Inductances in series and parallel.**

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

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

MUTUAL INDUCTANCE

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

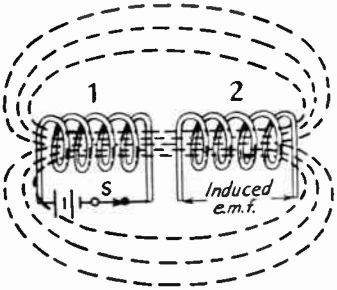


Fig. 2-15 — **Mutual inductance. When the switch, *S*, is closed current flows through coil No. 1, setting up a magnetic field that induces an e.m.f. in the turns of coil No. 2.**

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.

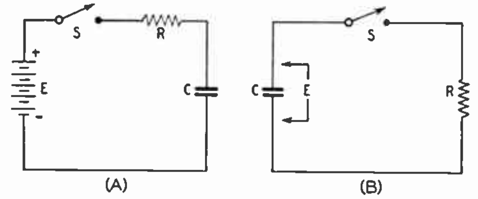


Fig. 2-16 — Schematics illustrating the time constant of an *RC* circuit.

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

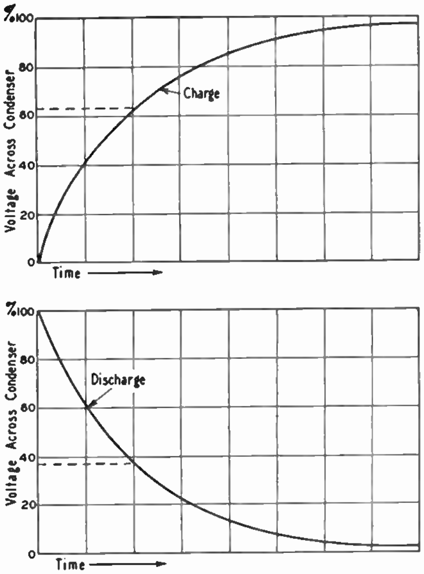


Fig. 2-17 — **How the voltage across a condenser rises, with time, when a condenser is charged through a resistor. The lower curve shows the way in which the voltage decreases across the condenser terminals on discharging through the same resistor.**

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,

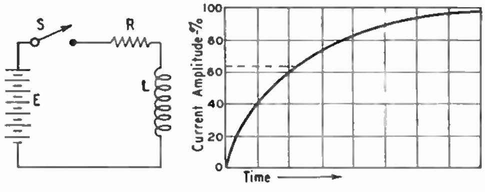


Fig. 2-18 — Time constant of an *LR* circuit.

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

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.

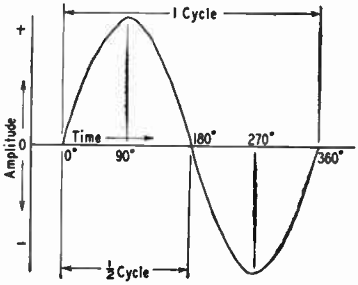


Fig. 2-19 — An a.c. cycle is divided off into 360 degrees that are used as a measure of time or phase.

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.

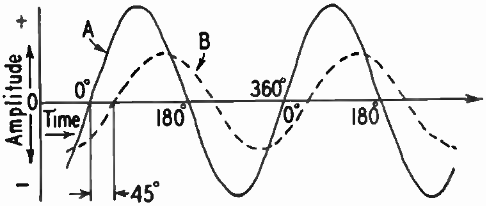


Fig. 2-20 — When two waves of the same frequency start their cycles at slightly different times, the time difference or phase difference is measured in degrees. In this drawing wave *B* starts 45 degrees (one-eighth cycle) later than wave *A*, and so lags 45 degrees behind *A*.

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.

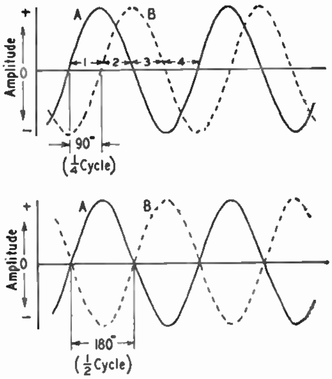


Fig. 2-21 — Two important special cases of phase difference. In the upper drawing, the phase difference between *A* and *B* is 90 degrees; in the lower drawing the phase difference is 180 degrees.

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.*

Capacitive Reactance

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

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

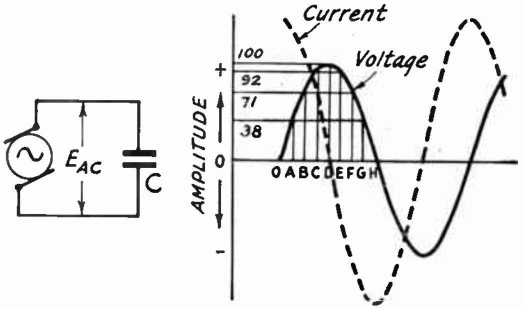


Fig. 2-22 — Voltage and current phase relationships when an alternating voltage is applied to a condenser.

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

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

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.

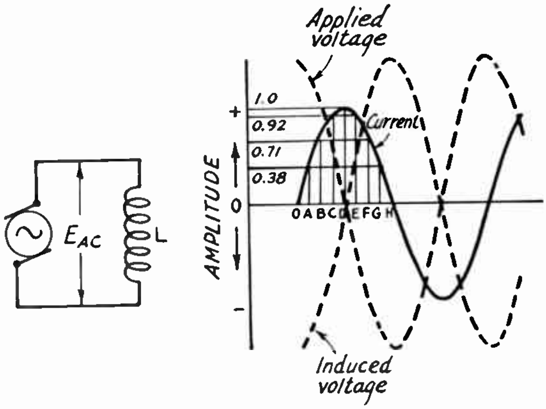


Fig. 2-23 — Phase relationships between voltage and current when an alternating voltage is applied to an inductance.

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

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

Ohm's Law for Reactance

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

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

$$E = IX$$

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

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

The reactance may be either inductive or capacitive.

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

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

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

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

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 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.

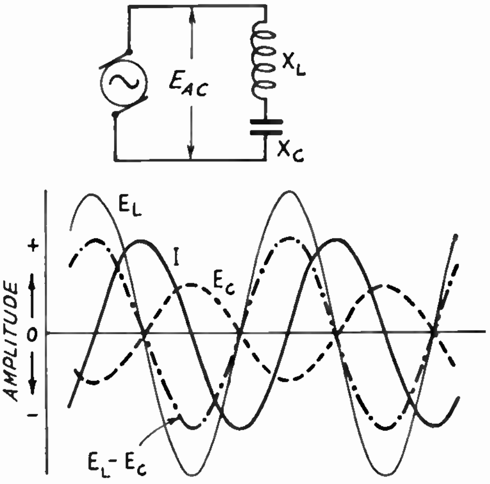


Fig. 2-24 — Current and voltages in a circuit having inductive and capacitive reactances in series.

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.

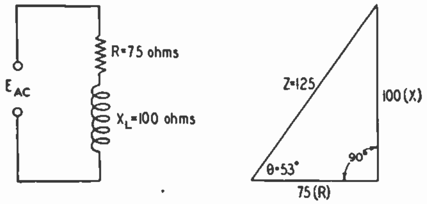


Fig. 2-25 — Resistance and inductive reactance connected in series.

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

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

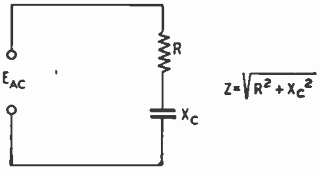
and the phase angle again is the angle whose tangent is equal to X/R. It must be remembered, however, that in this case the current *leads* the applied e.m.f., while in the resistance-inductance case it *lags* behind the voltage.

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.

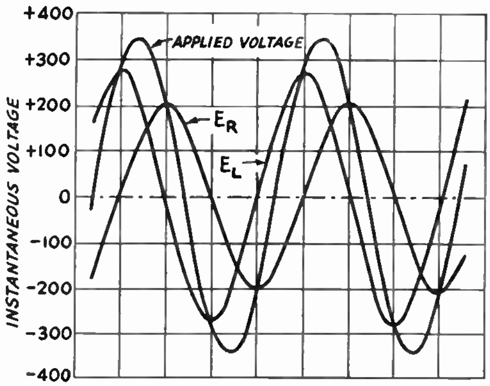


Fig. 2-27 — Voltage drops around the circuit of Fig. 2-25. Because of the phase relationships, the applied voltage is less than the arithmetical sum of the drops across the resistor and inductor.

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

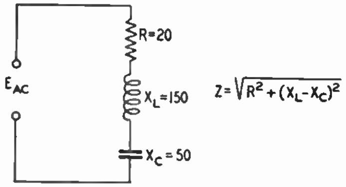


Fig. 2-28 — Resistance, inductive reactance, and capacitive reactance in series.

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

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

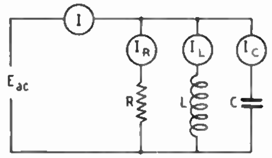


Fig. 2-29 — **Resistance, inductance and capacitance in parallel. Instruments connected as shown will read the total current, *I*, and the individual currents in the three branches of the circuit.**

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

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

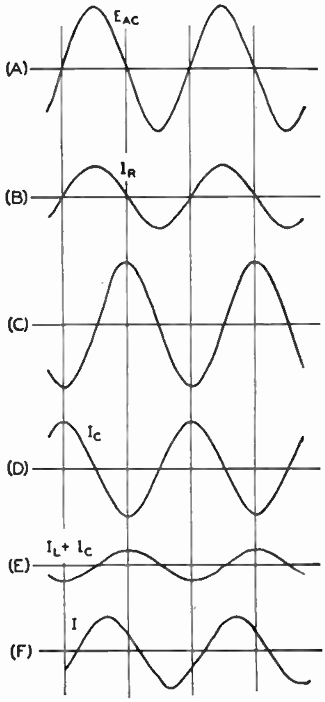


Fig. 2-30 — **Phase relationships between branch currents and applied voltage for the circuit of Fig. 2-29. The total current through *L* and *C* in parallel ($I_L + I_C$) and the total current in the entire circuit (*I*) also are shown.**

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-

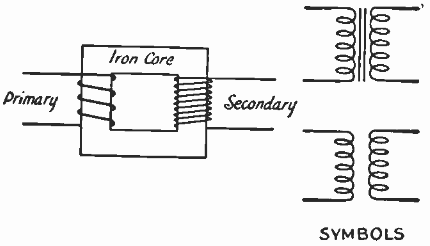


Fig. 2-31 — The transformer. Power is transferred from the primary coil to the secondary by means of the magnetic field. The upper symbol at right indicates an iron-core transformer, the lower one an air-core transformer.

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

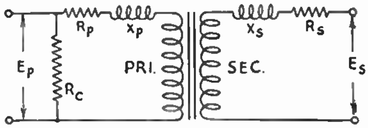


Fig. 2-32 — The equivalent circuit of a transformer includes the effects of leakage inductance and resistance of both primary and secondary windings. The resistance *Rc* is an equivalent resistance representing the constant core losses. Since these are comparatively small, their effect may be neglected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

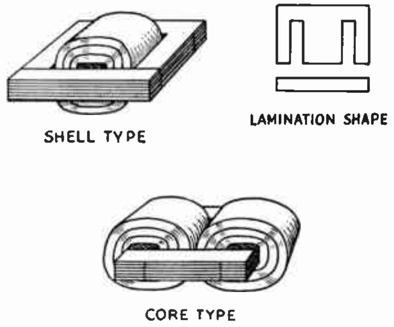


Fig. 2-33 — Two common types of transformer construction. Core pieces are interleaved to provide a continuous magnetic path with as low reluctance as possible.

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.

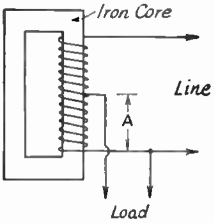


Fig. 2-34 — The autotransformer is based on the transformer principle, but uses only one winding. The line and load currents in the common winding (*A*) flow in opposite directions, so that the resultant current is the difference between them. The voltage across *A* is proportional to the turns ratio.

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 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.

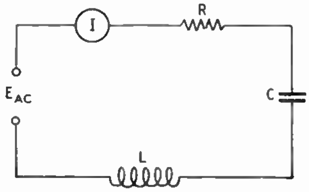


Fig. 2-35 — A series circuit containing L, C and R is "resonant" at the applied frequency when the reactance of C is equal to the reactance of L.

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

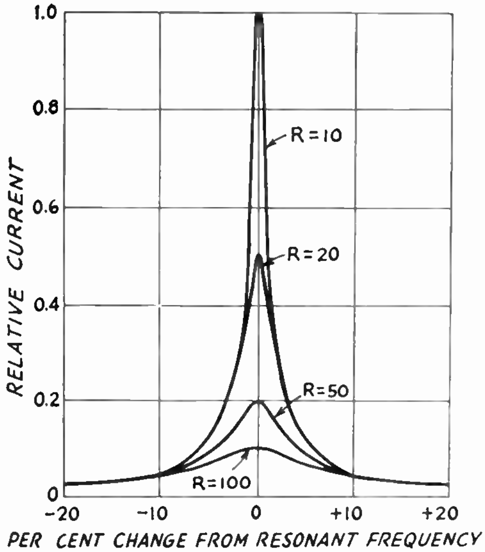


Fig. 2-36 — **Current in a series-resonant circuit with various values of series resistance. The values are arbitrary and would not apply to all circuits, but represent a typical case. It is assumed that the reactances (at the resonant frequency) are 1000 ohms (minimum Q = 10). Note that at frequencies 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.

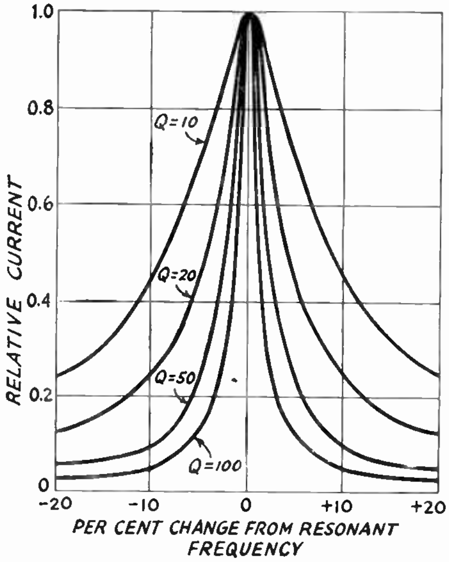


Fig. 2-37 — **Current in series-resonant circuits having different Qs. In this graph the current at resonance is assumed to be the same in all cases. The lower the Q, the more slowly the current decreases as the applied frequency is moved away from resonance.**

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 smaller and that of C higher at low frequencies; there is only partial cancellation of the two reactive currents and the line current therefore is larger than the current taken by R alone. At frequencies ***above*** resonance the situation is reversed and more current flows through C than through L, so the line current again increases. The current at resonance, being determined wholly by R, will be small if R is large and large if R is small.

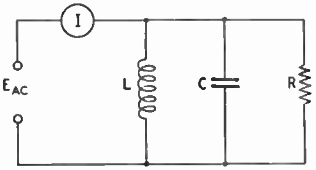


Fig. 2-38 — Circuit illustrating parallel resonance.

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

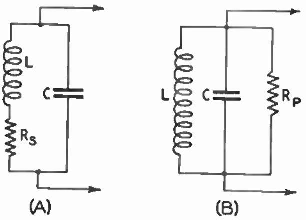


Fig. 2-39 — Series and parallel equivalents when the two circuits are resonant. The series resistor, R_s, in A can be replaced by an equivalent parallel resistor, R_p, in B, and vice versa.

Parallel and series resonant circuits are quite alike in some respects. For instance, the circuits given at A and B in Fig. 2-39 will behave identically, when an external voltage is applied, if (1) L and C are the same in both cases; and (2) R_p multiplied by R_s equals the square of the reactance (at resonance) of either L or C. When these conditions are met the two circuits will have the same 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

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

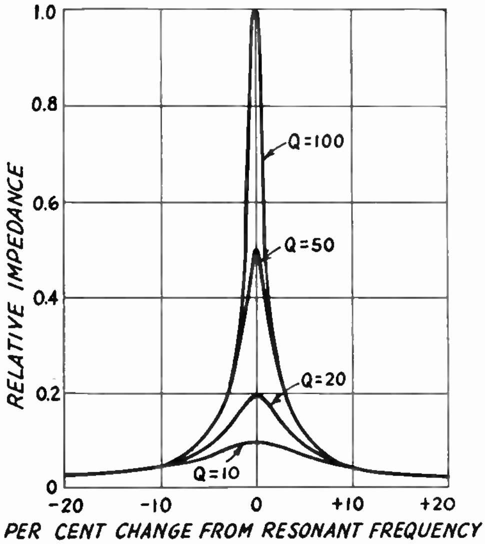


Fig. 2-40 — **Relative impedance of parallel-resonant circuits with different Qs. These curves are similar to those in Fig. 2-37 for current in a series-resonant circuit. The effect of Q on impedance is most marked near the resonant frequency.**

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

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

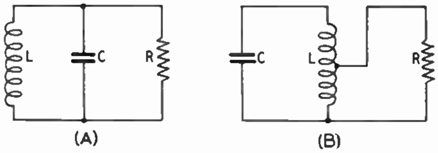


Fig. 2-41 — **The equivalent circuit of a resonant circuit delivering power to a load. The resistor R represents the load resistance. At B the load is tapped across part of L, which by transformer action is equivalent to using a higher load resistance across the whole circuit.**

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

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

L/C Ratio

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

LC Constants

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

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

where L = Inductance in microhenrys (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. $L = 1900/C = 1900/25 = 76\ \mu$h.
50 $\mu\mu$fd. $L = 1900/C = 1900/50 = 38\ \mu$h.
100 $\mu\mu$fd. $L = 1900/C = 1900/100 = 19\ \mu$h.
500 $\mu\mu$fd. $L = 1900/C = 1900/500 = 3.8\ \mu$h.

COUPLED CIRCUITS

Energy Transfer and Loading

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

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

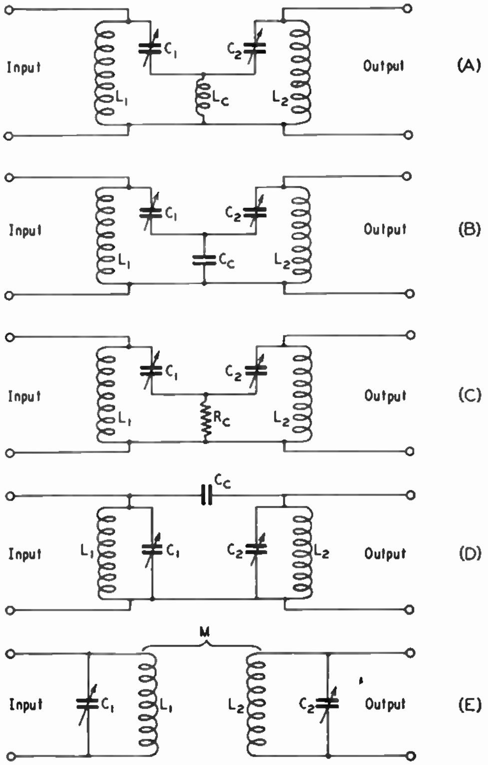


Fig. 2-42 — Basic methods of circuit coupling.

Coupling by a Common Circuit Element

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

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

Capacitive Coupling

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

Inductive Coupling

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

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

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

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

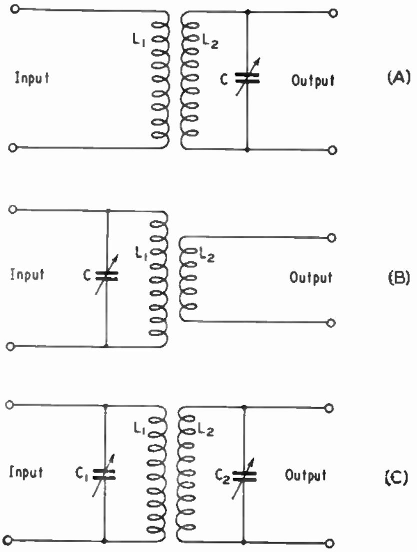


Fig. 2-43 — Types of inductively-coupled circuits. In A and B, one circuit is tuned, the other untuned. C shows the method of coupling between two tuned circuits.

Coupled Resonant Circuits

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

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

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

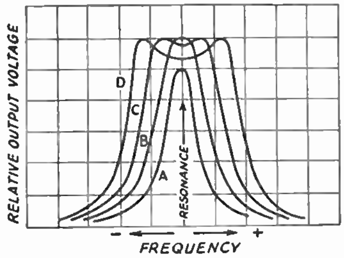


Fig. 2-44 — Showing the effect on the output voltage from the secondary circuit of changing the coefficient of coupling between two resonant circuits independently tuned to the same frequency. The voltage applied to the primary is held constant in amplitude while the frequency is varied, and the output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

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

Selectivity

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

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

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

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

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

Link Coupling

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

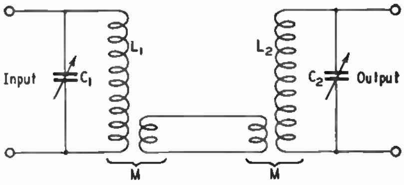


Fig. 2-45 — Link coupling. The mutual inductances at both ends of the link are equivalent to mutual inductance between the tuned circuits, and serve the same purpose.

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

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.

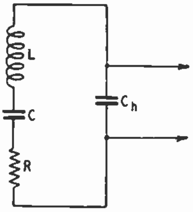


Fig. 2-46 — Equivalent circuit of a crystal resonator. L, C and R are the electrical equivalents of mechanical properties of the crystal; C_h is the capacitance of the electrodes with the crystal plate between them.

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

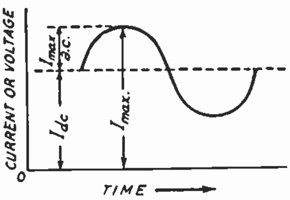


Fig. 2-47 — Pulsating current, composed of an alternating current or voltage superimposed on a steady direct current or voltage.

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.

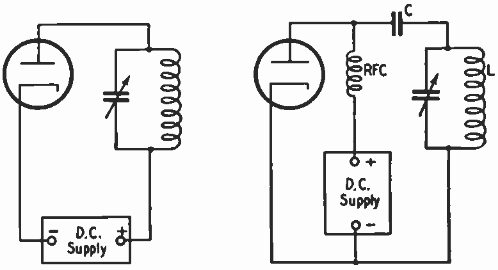
In the circuit at the right the direct current does not flow through the r.f. tuned circuit, but instead goes to the tube through a second coil, *RFC* **(radio-frequency choke).** Direct current cannot flow through *L* because a **blocking condenser,** *C*, is placed in the circuit to prevent it. (Without *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.



SERIES FEED PARALLEL FEED

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.

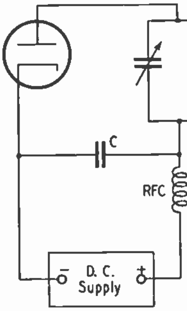


Fig. 2-49 — Typical use of a by-pass condenser in a series-feed circuit.

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 circuits to carry the audio frequencies around a d.c. supply.

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 a condenser 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 thus 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. Thus 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.

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 radio circuit. "Ground potential" means that there is no "difference of potential" — that is, no voltage — between the circuit point and the earth.

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* is connected to ground, so that the circuit has two ends each at the same voltage "above" ground.

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 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 at ground potential.

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. A metallic plate, called a **baffle shield,** inserted between two components also may suffice to prevent electrostatic coupling between them. It should be large enough to make the components invisible to each other.

Similar metallic shielding is used at radio frequencies to prevent magnetic coupling. The shielding effect increases with frequency and with the conductivity and thickness of the shielding material.

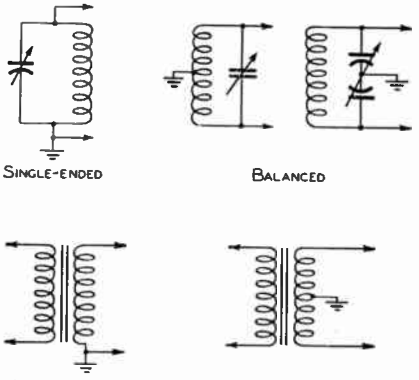


Fig. 2-50 — Single-ended and balanced circuits.

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, and between, the 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. For good magnetic shielding at audio frequencies it is necessary to enclose the coil in a container of high-permeability iron or steel. In this case the shield can be quite close to the coil without harming its performance.

Modulation, Heterodyning and Beats

Since one of the most widespread uses of radio frequencies is the wireless transmission of speech and music, it would be very convenient if the audio spectrum to be transmitted could simply be shifted up to some radio frequency, transmitted as radio waves, and shifted back down to the audio spectrum at the receiving point. Unfortunately, no simple method for doing this directly has ever been devised, although the *effect* is obtained and used in some advanced communications techniques.

Suppose the audio signal to be transmitted by radio is a pure 1000-cycle tone, and we wish to transmit it at some frequency around 1 Mc. (1,000,000 cycles). Simply adding the two frequencies (1000 and 1,000,000 cycles) in the same circuit doesn't necessarily do the trick — what does happen will depend upon the type of circuit in use. If the circuit or device obeys Ohm's Law at all values of applied voltage the two frequencies will simply be superimposed, and the resultant voltage across the circuit, or the current through it, will appear as in Fig. 2-51C. The amplitude at any instant is the sum of the amplitudes of the 1000- and 1,000,000-cycle signals at that instant. The r.f. in the circuit is not modified in any way by the presence of the audio signal, and thus any receiving equipment at a distance cannot possibly determine the nature of the audio signal.

There are, however, circuits and devices where Ohm's Law does not hold for all values of current or voltage. In such a circuit it is possible to have one frequency control the passage of the other. If, for example, a 1000-cycle tone is used to control the passage through the circuit of a 1-Mc. signal, the maximum r.f. output will be obtained when the 1000-cycle signal is at one peak and the minimum will occur at the other peak. The process is called **amplitude modulation**, and the effect is shown in Fig. 2-51D. Because the r.f. output varies at the modulation rate (1000 cycles), receiving equipment can detect these changes in amplitude and thus tell what the audio signal is.

It might be assumed that the only radio frequency present in such a signal is the original 1,000,000 cycles, but such is not the case. It will be found that two new frequencies have appeared. These are the sum (1,000,000 + 1000) and difference (1,000,000 − 1000) frequencies, and hence the radio frequencies appearing in the circuit after modulation are 999,000, 1,000,000 and 1,001,000 cycles.

Many circuits have been devised for obtaining amplitude modulation, and they will be treated in detail in later chapters. When an audio frequency is used to control the amplitude of a radio frequency, the process is generally called "amplitude modulation," as mentioned previously, but when a radio frequency modulates another radio frequency it is called **heterodyning.** However, they represent identical processes: the combination of signals in circuits that do not follow Ohm's Law. A general term for the sum and difference frequencies generated during heterodyning or amplitude modulation is **"beat frequencies,"** and a more specific one is **upper side frequency,** for the sum frequency, and **lower side frequency** for the difference frequency.

In the simple example, the modulating signal was assumed to be a pure tone, but the modulating signal can just as well be a *band* of frequencies making up speech or music. In this case, the side frequencies are grouped into what are called the **upper sideband** and the **lower sideband.** In any case, the frequency that is modulated is called the **carrier frequency.**

In A, B, C and D of Fig. 2-51, the sketches are obtained by plotting amplitude against time. However, it is equally helpful to be able to visualize the spectrum, or what a plot of amplitude *vs.* frequency looks like, at any given instant of time. E, F, G and H of Fig. 2-51 show the signals of Fig. 2-51A, B, C and D on an amplitude-*vs.*-frequency basis. Any one frequency is, of course, represented by a vertical line. Fig. 2-51H shows the side frequencies appearing as a result of the modulation process.

Since the carrier and the sidebands are radio frequencies, they can be radiated from the transmitting station and received at a distant point. If they are now passed through a suitable circuit that does not follow Ohm's Law, the sidebands will heterodyne with the carrier to give sum and difference frequencies. The sum frequencies are not used and are filtered out by tuned circuits, but the difference frequency is the original modulating signal. The process is called **demodulation** or **detection.**

Amplitude modulation (AM) is not the only possible type nor is it the only one in use. The *frequency* or the *phase* of the carrier can be varied in accordance with the modulating frequency or frequencies, to give **frequency modulation** (FM) or **phase modulation** (PM). (For practical circuits, see later chapters.) Here again, upper and lower sidebands are generated as a result of the modulation process, but instead of taking the simple form of the sum and difference frequencies they include higher-order sidebands as well. The signals are demodulated in frequency- or phase-sensitive circuits, as described later.

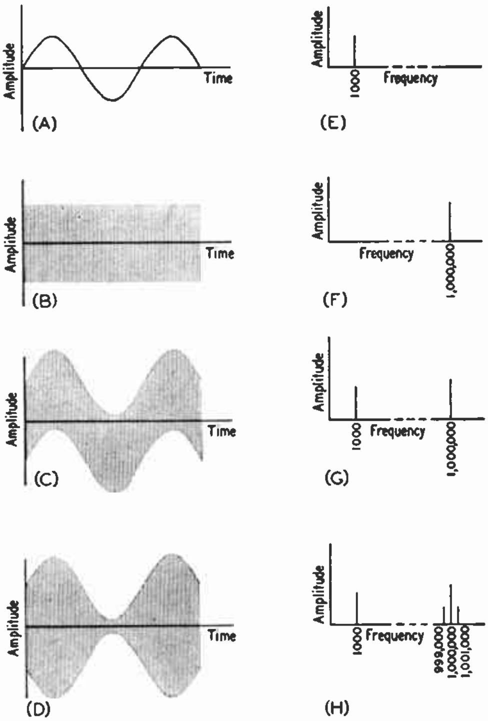


Fig. 2-51 — Amplitude-*vs.*-time and amplitude-*vs.*-frequency plots of various signals. (A) 1½ cycles of a 1000-cycle signal. (B) A 1,000,000-cycle signal plotted to the same scale as A. Because there are 1500 cycles during this time, they cannot be shown accurately. (C) The signals of A and B added in a circuit that obeys Ohm's Law. Although the limits vary above and below the horizontal axis, the amplitude of the 1,000,000-cycle signal is constant. (D) The signals of A and B combined in a circuit where Ohm's Law does not hold. The 1,000,000-cycle signal is *modulated* by the 1000-cycle signal, and the amplitude of the 1,000,000-cycle signal is not constant. (E), (F), (G), (H) Amplitude-*vs.*-frequency plots of the signals in A, B, C and D.

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 earlier.

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 an earlier section 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.

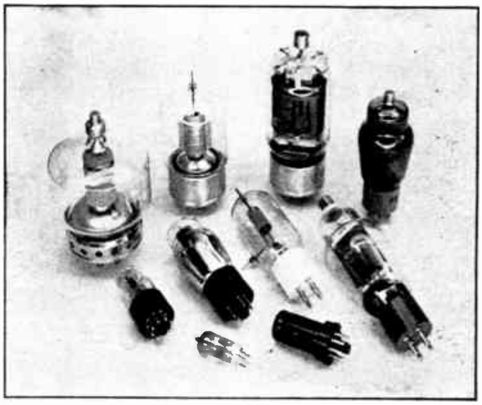
The most practical way to introduce a sufficiently-large number of electrons into the evacuated space is by **thermionic emission.**

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.

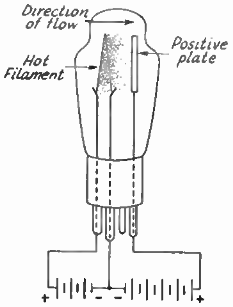


Fig. 3-1 — Conduction by thermionic emission in a vacuum tube. One battery is used to heat the filament to a temperature that will cause it to emit electrons. The other battery makes the plate positive with respect to the filament, thereby causing the emitted electrons to be attracted to the plate. Electrons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

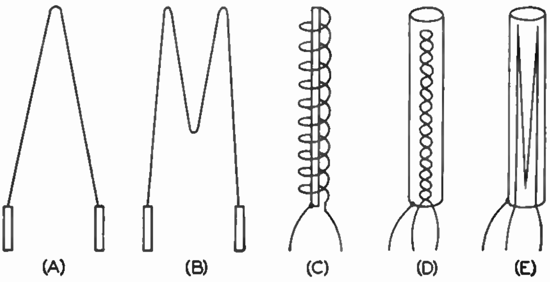


Fig. 3-2 — Types of cathode construction. Directly-heated cathodes or filaments are shown at A, B, and C. The inverted V filament is used in small receiving tubes, the M in both receiving and transmitting tubes. The spiral filament is a transmitting-tube type. The indirectly-heated cathodes at D and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

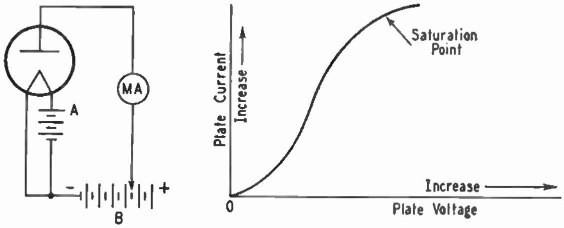


Fig. 3-3 — The diode, or two-element tube, and a typical curve showing how the plate current depends upon the voltage applied to the plate.

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.

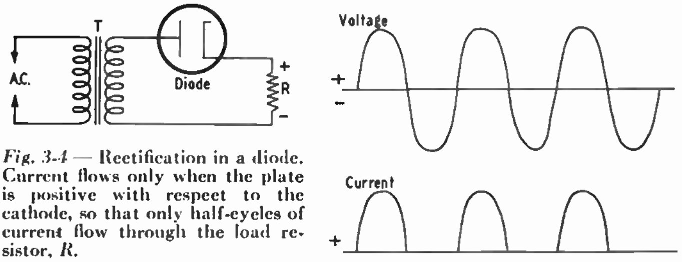


Fig. 3-4 — Rectification in a diode. Current flows only when the plate is positive with respect to the cathode, so that only half-cycles of current flow through the load resistor, *R*.

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

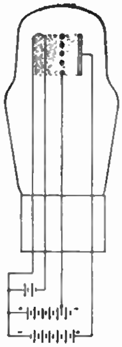


Fig. 3-5 — Construction of an elementary triode vacuum tube, showing the filament, grid (with an end view of the grid wires) and plate. The relative density of the space charge is indicated roughly by the dot density.

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.

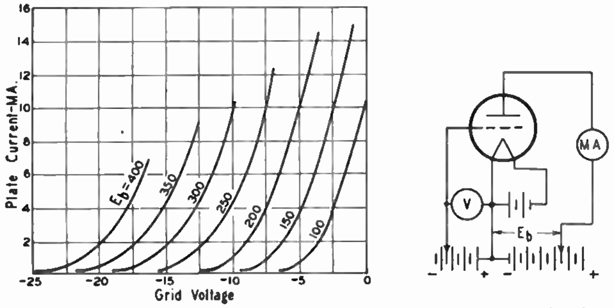


Fig. 3-6 — Grid-voltage-*vs.*-plate-current curves at various fixed values of plate voltage (E_b) for a typical small triode. Characteristic curves of this type can be taken by varying the battery voltages in the circuit at the right.

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 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 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.

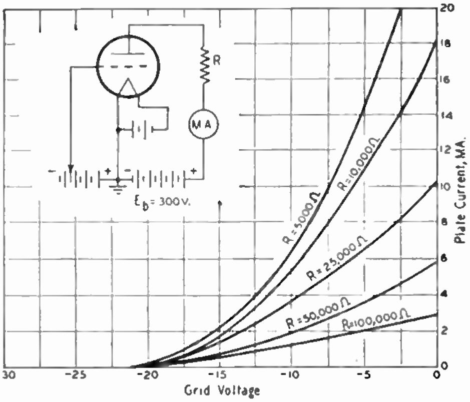


Fig. 3-7 — **Dynamic characteristics of a small triode with various load resistances from 5000 to 100,000 ohms.**

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

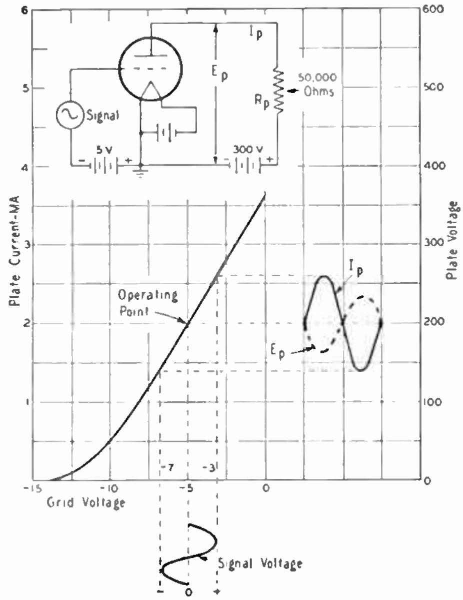


Fig. 3-8 — Amplifier operation. When the plate current varies in response to the signal applied to the grid, a varying voltage drop appears across the load, R_p, as shown by the dashed curve, E_p. I_p is the plate current.

graph. When the plate current is maximum, the instantaneous voltage drop in R_p is 50,000 × 0.00265 = 132.5 volts; when the plate current is minimum the instantaneous voltage drop in R_p is 50,000 × 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

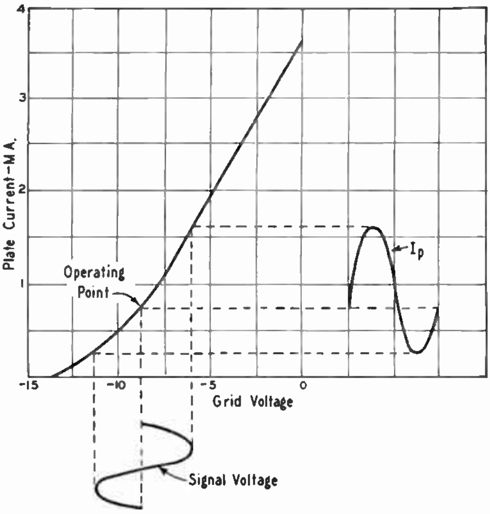


Fig. 3-9 — Harmonic distortion resulting from choice of an operating point on the curved part of the tube characteristic. The lower half-cycle of plate current does not have the same shape as the upper half-cycle.

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 an earlier chapter, 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.

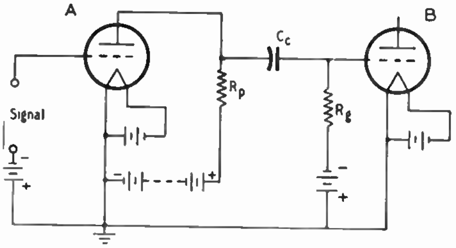
Amplifier Output Circuits

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.

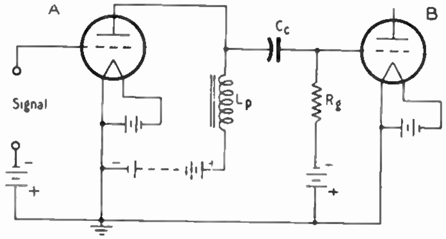
Three types of coupling are in common use at audio frequencies. These are **resistance coupling, impedance coupling, and transformer coupling.** They are shown in Fig. 3-10. In all three cases the output is shown coupled to the grid circuit of a subsequent amplifier tube, but the same types of circuits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

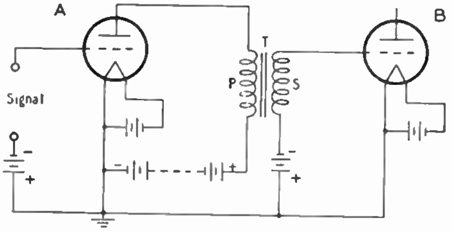
The grid resistor, R_g, usually has a rather high value (0.5 to 2 megohms). The reactance of the coupling condenser, C_c, must be low enough compared to the resistance of R_g so that the a.c. voltage drop in C_c is negligible at the lowest frequency to be amplified. If R_g is at least 0.5 megohm, a 0.1-μfd.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



Resistance Coupling



Impedance Coupling



Transformer Coupling

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

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

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

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

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

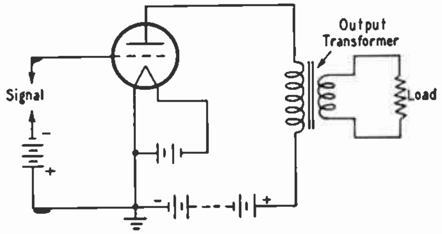


Fig. 3-11 — An elementary power-amplifier circuit in which the power-consuming load is coupled to the plate circuit through an impedance-matching transformer.

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

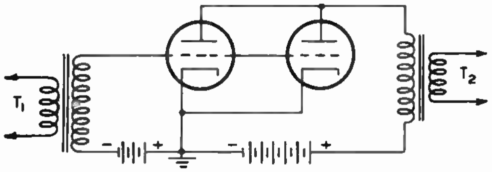
The end result of any amplification is that the amplified signal does some *work*. As a familiar example, an audio-frequency amplifier usually drives a loudspeaker that in turn produces sound waves. The greater the amount of a.f. *power* supplied to the 'speaker, the louder the sound it will produce.

In some amplifiers, therefore, *power* output rather than voltage is the primary consideration. It was mentioned in an earlier chapter 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.

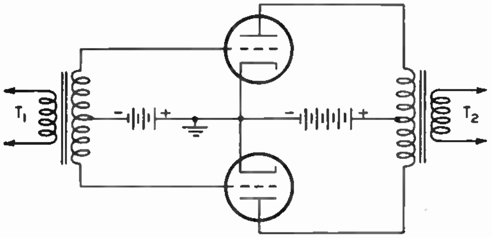
Fig. 3-11 shows an elementary power-amplifier circuit. It is simply a transformer-coupled amplifier with the load connected to the secondary. Although the load is shown as a resistor, it actually would be some device, such as a loudspeaker, that employs the power usefully. The resistance of the actual 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



PARALLEL



PUSH-PULL

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.

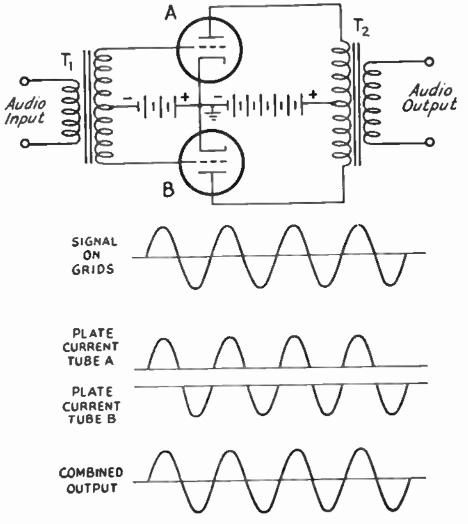


Fig. 3-13 — Class B amplifier operation.

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 later chapters. 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

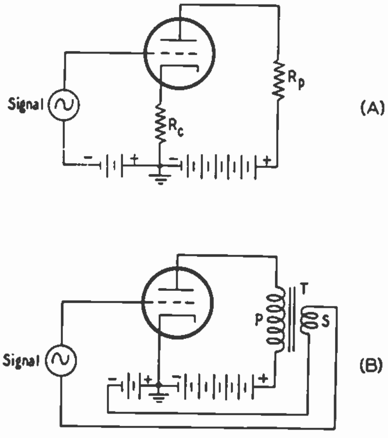


Fig. 3-14 — Circuits for producing feed-back. In A, part of the a.c. plate voltage appears across the cathode resistor, R_c, and is therefore also applied between grid and cathode. The feed-back is negative in this case. In B, the voltage that is generated in the secondary of the transformer is inserted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

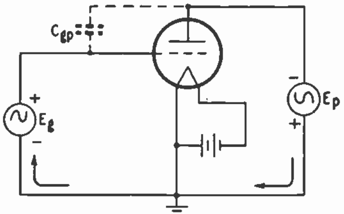


Fig. 3-15 — The a.c. voltage appearing between the grid and plate of the amplifier is the sum of the signal voltage and the output voltage, as shown by this simplified circuit. Instantaneous polarities are indicated.

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.

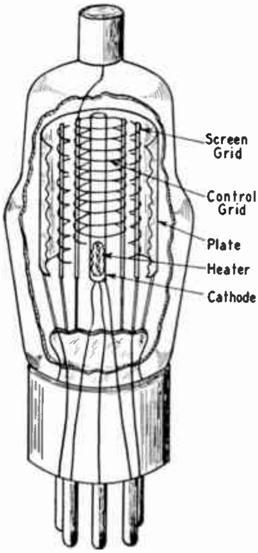


Fig. 3-16 — Representative arrangement of elements in a screen-grid tube, with front part of plate and screen grid cut away. In this drawing the control-grid connection is made through a cap on the top of the tube, thus eliminating the capacitance that would exist between the plate- and grid-lead wires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

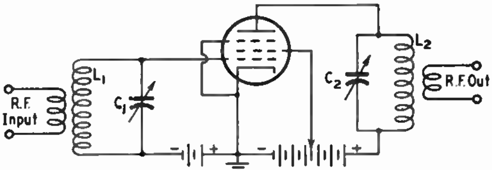


Fig. 3-17 — Simplified pentode r.f.-amplifier circuit. L_1C_1 and L_2C_2 are tuned to the same frequency.

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.

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.

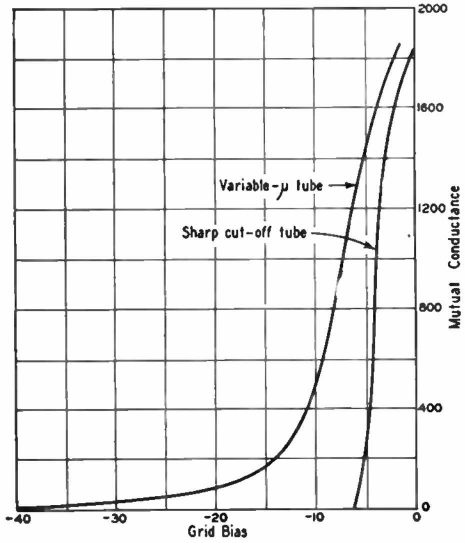


Fig. 3-18 — Curves showing the relationship between mutual conductance and negative grid bias for two small receiving pentodes, one a sharp cut-off type and the other a variable-μ type.

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 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.

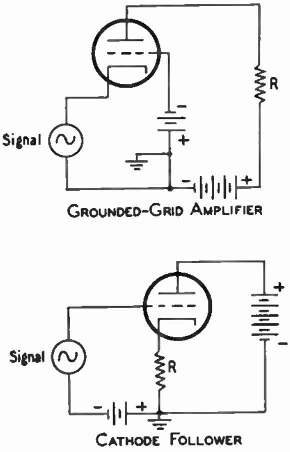


Fig. 3-19 — In the upper circuit, the grid is the junction point between the input and output circuits. In the lower drawing, the plate is the junction. In either case the output is developed in the load resistor, *R*, and may be coupled to a following amplifier by the usual methods.

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.

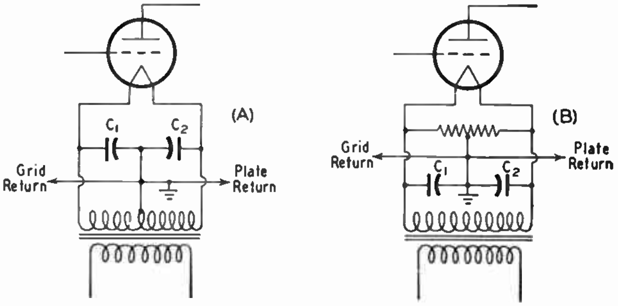


Fig. 3-20 — Filament center-tapping methods for use with directly-heated tubes.

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 μμfd. 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.

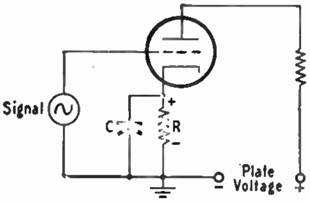


Fig. 3-21 — Cathode biasing. *R* is the cathode resistor and *C* is the cathode by-pass condenser.

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 feedback is positive) or loss of gain (if the feedback 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 a later chapter, 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 the chapter on Power Supplies.

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.

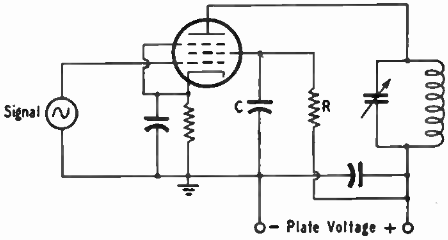


Fig. 3-22 — Screen-voltage supply for a pentode tube through a dropping resistor, *R*. The screen by-pass condenser, *C*, must have low enough reactance to bring the screen to ground potential for the frequency or frequencies being amplified.

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.

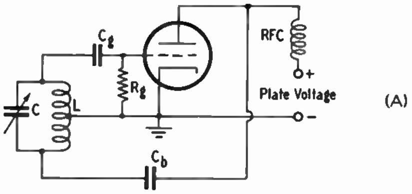
The *amount* of feed-back depends on the *position* of the tap. If the tap is too close to either end of the coil the circuit will not oscillate. If the tap is too near the grid end the voltage drop is too small to give enough feed-back, and if it is too near the plate end the impedance between the cathode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

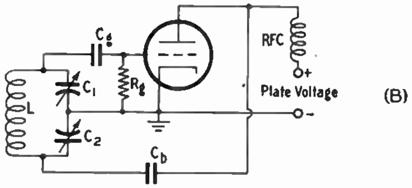
Condenser C_g is the **grid condenser.** It and R_g (the **grid leak**) are used for the purpose of obtaining grid bias for the tube. In this (and practically all) oscillator circuits the tube generates its own bias. When the grid end of the tuned circuit is positive with respect to the cathode, the 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.

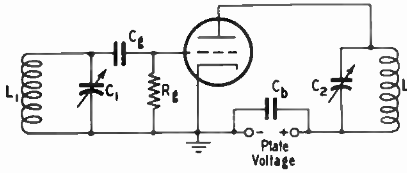


Fig. 3-24 — The tuned-plate tuned-grid oscillator.

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.

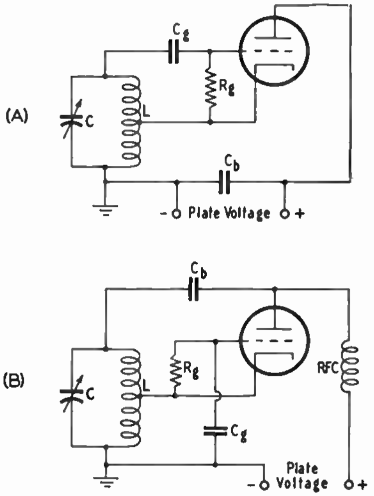


Fig. 3-25 — Showing how the r.f. ground on a typical oscillator circuit (Hartley) may be placed on either the plate (A) or grid (B) instead of the more conventional method of grounding the cathode. Provided the proper provisions are made for supplying cathode and plate voltages, the circuit operation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

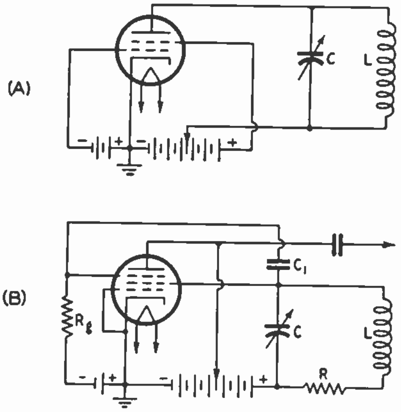


Fig. 3-26 — Negative-resistance oscillator circuits. A, dynatron; B, transitron.

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 amateurs have in the past. In fact, it is through amateur efforts that most of the extended-range possibilities of various radio frequencies have been discovered, either through accident or long and careful investigation.

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.

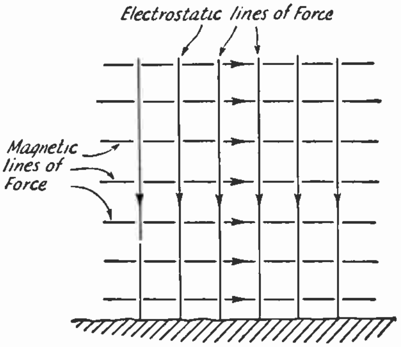


Fig. 4-1 — Representation of electrostatic and electromagnetic lines of force in a radio wave. Arrows indicate instantaneous directions of the fields for a wave traveling toward the reader. Reversing the direction of one set of lines would reverse the direction of travel.

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.

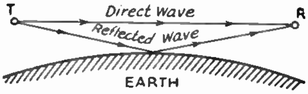


Fig. 4-2 — Showing how both direct and reflected waves may be received simultaneously in v.h.f. transmission.

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.

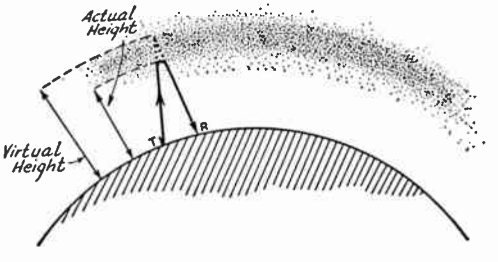


Fig. 4-3 — Bending in the ionosphere, and the echo or reflection method of determining virtual height.

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. 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** may be observed on any frequency, depending upon the conditions, and it is always characterized by a flutter on all signals that 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. Abnormally short distances for the frequency may be covered if the ionized cloud is situated midway between transmitter and receiver, or is of any 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 is readily determined by the reception of 14-Mc. signals from stations 500 to 100 miles away, or by the reception of 28-Mc. signals from distances of 1000 to 300 miles. Over these abnormally short paths, the signal strengths may often rise to high intensities, of an order usually associated only with lower-frequency transmissions.

Scatter

Scatter signals are heard on any band, but are more easily recognizable on the higher frequencies because of the extended skip zone. They are signals reflected from large discontinuities at a distance, such as sharp concentrations of ionization in any of the normal layers, sporadic-*E* clouds or (rarely) large land objects. They result in one's hearing signals within the normal skip zone. Scatter signals are never very loud, and have a slight flutter characteristic. A further indication of scatter reflection is that, when beam antennas are used to indicate the direction of arrival of the wave, the ray path is not necessarily the direct route but can even be at right angles or in the opposite direction.

Meteor Trails

Another phenomenon generally encountered in the 28-Mc. band, but also observed in the 14- and 50-Mc. bands, is one characterized by sudden bursts of intensity of a signal. These bursts last less than a second, generally, and are caused by reinforced reflection from the ionized trail of a meteor. The meteor, entering the earth's atmosphere at high velocity, heats by friction against the atmosphere and leaves a trail of ionized atmosphere. It takes a finite time for the ionized molecules to recombine, and during this time a small ionized cloud exists. If it is in the ray path of a signal, it may serve to reinforce the signal and cause the burst in intensity. When the meteor is moving in a direction somewhat parallel to the ray path, it can induce a rising or falling "whistle" on the signal, for a second or so. The effects of bursts and whistles can be observed at any time during the day or night, if there is any marked meteor activity, and during rare "meteor showers" the ionized clouds can serve in almost the same manner that sporadic-*E* does to make long-distance work possible on 50 Mc.

Wave Angle

The smaller the angle at which a wave leaves the earth, the less will be the bending required in the ionosphere to bring it back and, in general, the greater the distance between the point where it leaves the earth and that at which it returns. This is shown in Fig. 4-4. The vertical angle which the wave makes with a tangent to the earth is called the **wave angle or angle of radiation.**

Skip Distance

Since greater bending is required to return the wave to earth when the wave angle is high, at the higher frequencies the refraction frequently is not enough to give the required bending unless the wave angle is smaller than a certain angle called the **critical angle.** This is illustrated in Fig. 4-4, where waves at angles of *A* or less give useful signals while waves sent at higher angles penetrate the layer and are not returned. The distance between *T* and R_1 is, therefore, the shortest possible distance over which communication by normal ionospheric refraction can be accomplished.

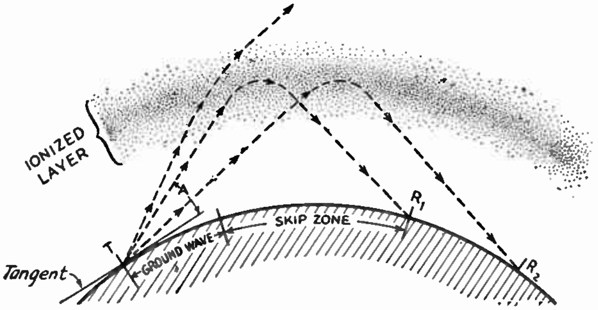


Fig. 4-4 — Refraction of sky waves, showing the critical wave angle and the skip zone. Waves leaving the transmitter at angles above the critical (greater than *A*) are not bent enough to be returned to earth. As the angle is increased, the waves return to earth at increasingly greater distances.

The area between the end of the useful ground wave and the beginning of ionospheric-wave reception is called the **skip zone.** The extent of skip zone depends upon the frequency and the state of the ionosphere, and is greater the higher the transmitting frequency and the lower the critical frequency. Skip distance depends also upon the height of the layer in which the refraction takes place, the higher layers giving longer skip distances for the same wave angle. Wave angles at the transmitting and receiving points are usually, although not always, approximately the same for any given wave path.

It is readily possible for the ionospheric wave to pass through the *E* layer and be refracted back to earth from the *F*, F_1 or F_2 layers. This is because the critical frequencies are higher in the latter layers, so that a signal too high in frequency to be returned by the *E* layer can still come back from one of the others, depending upon the time of day and the existing conditions. Depending upon the wave angle and the frequency, it is sometimes possible to carry on communication via either the *E* or F_1-F_2 layers on the same frequency.

Multihop Transmission

On returning to the earth the wave can be reflected upward and travel again to the ionosphere. There it may once more be refracted, and again bent back to earth. This process may be repeated several times. Multihop propagation of this nature is necessary for transmission over great distances because of the limited heights of the layers and the curvature of the earth, since at the lowest useful wave angles (of the order of a few degrees, waves at lower angles generally being absorbed rapidly at high frequencies by being in contact with the earth) the maximum one-hop distance is about 1250 miles for refraction from the *E* layer and around 2500 miles for the F_2 layer. However, ground losses absorb some of the energy from the wave on each reflection (the amount of the loss varying with the type of ground and being least for reflection from sea water). Thus, when the distance permits, it is better to have one hop rather than several, since the multiple reflections introduce losses that are higher than those caused by the ionosphere alone.

Fading

Two or more parts of the wave may follow slightly different paths in traveling to the receiving point, in which case the difference in path lengths will cause a phase difference to exist between the wave components at the receiving antenna. The field strength therefore may have any value between the numerical sum of the components (when they are all in phase) and zero (when there are only two components and they are exactly out of phase). Since the paths change from time to time, this causes a variation in signal strength called fading. Fading can also result from the combination of single-hop and multihop waves, or the combination of a ground wave with an ionospheric or tropospheric wave. Such a condition gives rise to an area of severe fading near the limiting distance of the ground wave, better reception being obtained at both shorter and longer distances where one component or the other is considerably stronger. Fading may be rapid or slow, the former type usually resulting from rapidly-changing conditions in the ionosphere, the latter occurring when transmission conditions are relatively stable.

It frequently happens that transmission conditions are different for waves of slightly different frequencies, so that in the case of voice-modulated transmission, involving sidebands differing slightly from the carrier in frequency, the carrier and various sideband components may not be propagated in the same relative amplitudes and phases they had at the transmitter. This effect, known as **selective fading,** causes severe distortion of the signal.

Tropospheric Propagation

Changes in refractive index of air masses in the lower atmosphere often permit work over greater-than-normal distances on 28 Mc. and higher frequencies. The effect can be observed on 28 Mc., but it is generally more marked on 50 and 144 Mc. The subject is treated in detail in a later chapter.

● 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 a result of their experience, not of 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. The only noise that is amplified is that which falls within the passband of the receiver, so 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.)

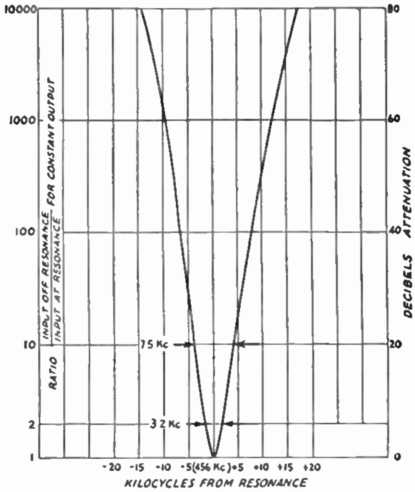


Fig. 5-1 — Typical selectivity curve of a modern superheterodyne receiver. Relative response is plotted against deviations above and below the resonance frequency. The scale at the left is in terms of voltage ratios, the corresponding decibel steps are shown at the right.

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 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.

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 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 (see "Modulation, Heterodyning and Beats"). 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 for a.m. is the diode. 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. In audio work the load resistor, R_1, is usually 0.1 megohm or higher, 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

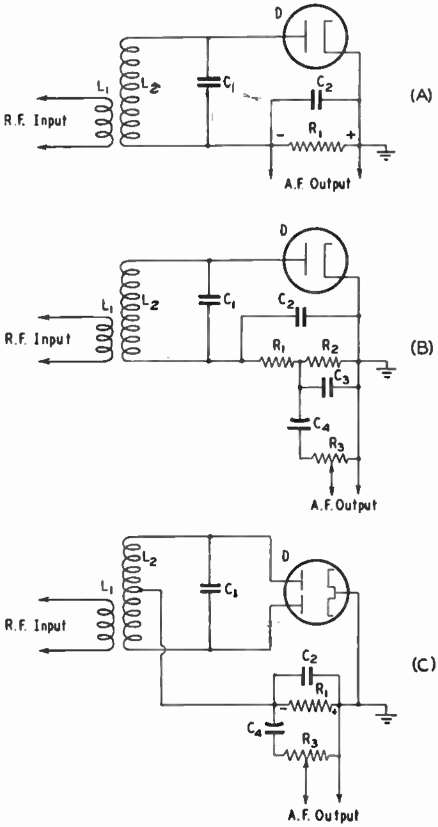


Fig. 5-2 — Simplified and practical diode detector circuits. A, the elementary half-wave diode detector; B, a practical circuit, with r.f. filtering and audio output coupling; C, full-wave diode detector, with output coupling indicated. The circuit, L_2C_1, is tuned to the signal frequency; typical values 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.

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.

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.

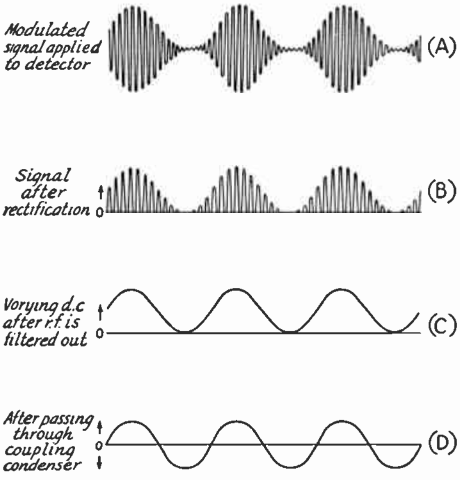


Fig. 5-3 — Diagrams showing the detection process.

Compared with other detectors, the sensitivity of the diode is low, normally running around 0.8 in audio work. Since the diode consumes power, the Q of the tuned circuit is reduced, bringing about a reduction in selectivity. The loading effect of the diode is close to one-half the load resistance. The detector linearity is good, and the signal-handling capability is high.

Plate Detectors

The plate detector is arranged so that rectification of the r.f. signal takes place in the plate circuit of the tube. 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. 5-4. C_3 is the plate by-pass condenser, and, with *RFC*, prevents r.f. from appearing in the output. R_1 is the cathode resistor which provides the operating grid bias, and C_2 is a by-pass for both radio and audio frequencies across R_1. R_2 is the plate load resistance across which a voltage appears as a result of the rectifiying action described above. C_4 is the output coupling condenser. In the pentode circuit at B, R_3 and R_4 form a voltage divider to supply the proper potential (about 30 volts) to the screen, and C_5 is a by-pass condenser between screen and cathode. C_5 must have low reactance for both radio and audio frequencies.

In general, transformer coupling from the plate circuit of a plate detector is not satisfactory, because the plate impedance even of a triode is very high when the bias is set near the plate-current cut-off point. Impedance coupling may be used in place of the resistance coupling shown in Fig. 5-4. Usually 100 henrys or more inductance is required.

The plate detector is more sensitive than the diode since there is some amplifying action in the tube. It will handle large signals, 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. 5-5 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.

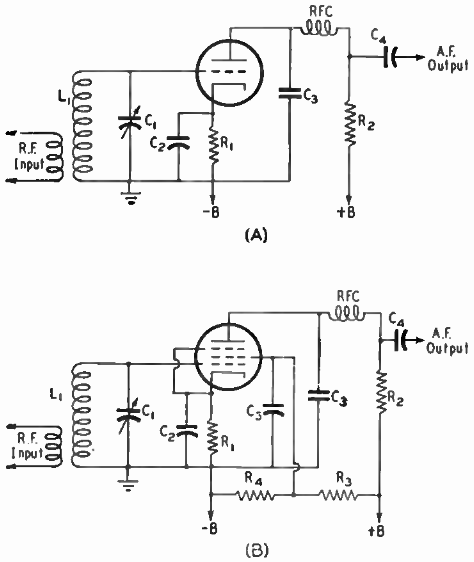


Fig. 5-4 — **Circuits for plate detection. A, triode; B, pentode. The input circuit, L_1C_1, is tuned to the signal frequency. Typical values for the other components are:**

Component	*Circuit A*	*Circuit B*
C_2	0.5 μfd. or larger.	0.5 μfd. or larger.
C_3	0.001 to 0.002 μfd.	250 to 500 μμfd.
C_4	0.1 μfd.	0.1 μfd.
C_5		0.5 μfd. or larger.
R_1	25,000 to 150,000 ohms.	10,000 to 20,000 ohms.
R_2	50,000 to 100,000 ohms.	100,000 to 250,000 ohms.
R_3		50,000 ohms.
R_4		20,000 ohms.
RFC	2.5 mh.	2.5 mh.

Plate voltages from 100 to 250 volts may be used. Effective screen voltage in B should be about 30 volts.

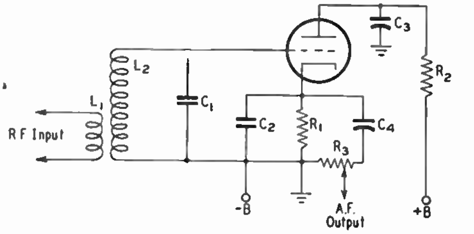


Fig. 5-5 — **The infinite-impedance detector. The input circuit, L_2C_1, is tuned to the signal frequency. Typical values for the other components are:**

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.

The grid-leak detector is a combination diode rectifier and audio-frequency amplifier. In the circuits of Fig. 5-6, the grid corresponds to the diode plate and the rectifying action is exactly the same as in a diode. 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, T_1, L_4 and L_3 are the plate load resistances, C_3 is a by-pass condenser and *RFC* an r.f. choke to eliminate r.f. in the output circuit. With a triode, the load is almost always an audio transformer.

Since audio amplification is added to rectification, the grid-leak detector has considerably greater sensitivity than the diode. The sensitivity is further increased by using a screen-grid tube instead of a triode, as at 5-6 B and C. 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_2 and R_3 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 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 audio choke 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 regenerative grid-leak detector is higher than that of any other type. However, the many disadvantages of the detector commend it for use only in the simplest receivers. 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 degree of antenna coupling is often critical.

The circuits in Fig. 5-6 are regenerative, the feedback being obtained by feeding some signal to the grid back from the plate circuit. 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.

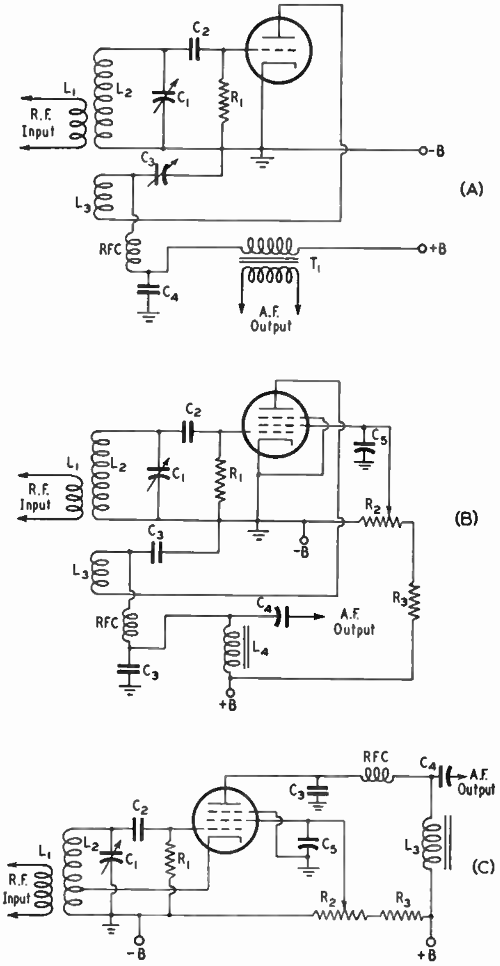


Fig. 5-6 — Triode and pentode regenerative detector circuits. The input circuit, L_2C_1, is tuned to the signal frequency. The grid condenser, C_2, should have a value of about 100 μμfd. in all circuits; the grid leak, R_1, may range in value from 1 to 5 megohms. The tickler coil, L_3, ordinarily will 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. 5-6 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 5-6B 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.

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. 5-6C is used, and then only at 14 Mc. and higher frequencies. Connecting one side of the heater supply to ground, or grounding the center-tap of the heater-transformer winding, is good practice to reduce hum, and the heater wiring should be kept as far as possible from the r.f. circuits.

House wiring, if of the "open" type, will have a rather extensive electrostatic field which may cause hum if the detector tube, grid lead, and grid condenser and leak are not 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. 5-7. It will be found that a

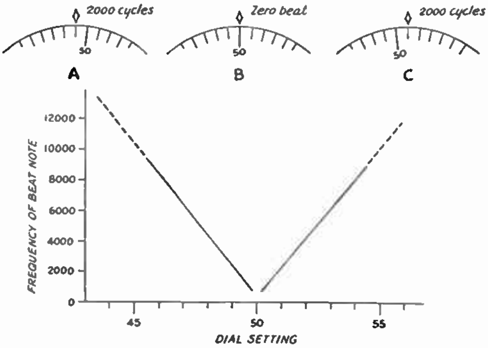


Fig. 5-7 — **As the tuning dial of a receiver is turned past a c.w. signal, the beat-note varies from a high tone down through "zero beat" (no audible frequency difference) and back up to a high tone, as shown at A, B and C. The curve is a graphical representation of the action. The beat exists past 8000 or 10,000 cycles but usually is not heard because of the limitations of the audio system.**

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 susceptible 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 coil and tuning condenser cannot be used for, say, 14 Mc. to 3.5 Mc., because of the impracticable maximum-to-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 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. The unused coils are sometimes short-circuited by the switch, to avoid the possibility of undesirable self resonances in the unused coils. It is not necessary if the coils are separated from each other by several coil diameters, or are mounted at right angles to each other.

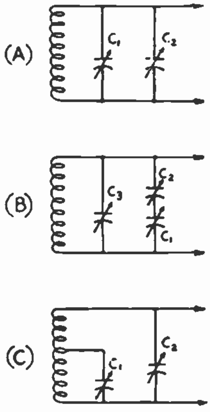


Fig. 5-8 — Essentials of the three basic band-spread tuning systems.

Another method is to use coils wound on forms with contacts (usually pins) which can be plugged in and removed from a socket. These coils are advantageous when space in a multiband receiver is at a premium. They are also very useful when considerable experimental work is involved, because they are easier to work on than coils clustered around a switch.

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. 5-8.

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 μμfd. 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 higher 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. 5-9, where C_1 is the trimmer and C_2 the tuning condenser. The use of the trimmer necessarily increases the minimum circuit capacity, but it is a necessity for satisfactory tracking. Midget condensers having maximum capacities of 15 to 30 μμfd. are commonly used.

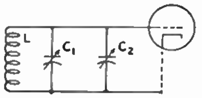


Fig. 5-9 — Showing the use of a trimmer condenser to set the minimum circuit capacity in order to obtain true tracking for gang-tuning.

The same methods are applied to bandspread circuits that must be tracked. The circuits are identical with those of Fig. 5-8. 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. 5-8B, and C_2 in Fig. 5-8C, 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 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. 5-10.

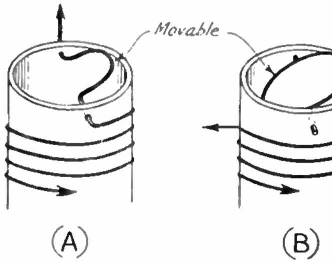


Fig. 5-10 — Methods of adjusting the inductance for ganging. The half-turn in A can be moved so that its magnetic field either aids or opposes the field of the coil. The shorted loop in B is not connected to the coil, but operates by induction. It will have no effect on the coil inductance when the axis 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.

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 undesired signal is called the **image.** It can cause unnecessary interference if it isn't eliminated.

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. 5-11. The variations are chiefly in the way in which the oscillator voltage is introduced. In 5-11A, 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 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. 5-11B, 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.

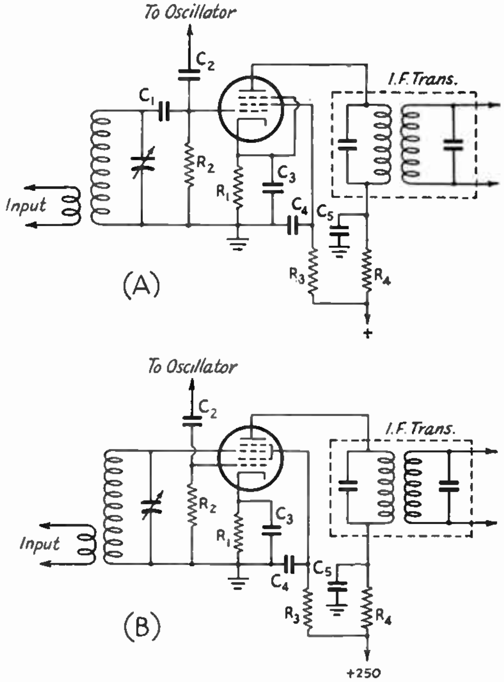


Fig. 5-11 — Typical circuits for separately-excited mixers. Grid injection of a pentode mixer is shown at A, and separate excitation of a pentagrid converter is given in B. Typical values for B will be found in Table 5-I — the values below are for the pentode mixer of A.

C_1 — 10 to 50 μμfd.
C_2 — 5 to 10 μμfd.
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.

Many receivers use pentagrid converters, and two typical circuits are shown in Fig. 5-12. The circuit shown in Fig. 5-12A, which is suitable for the 6K8 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. 5-12B can be used with a tube like the 6SA7, 7Q7, 6BA7 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.

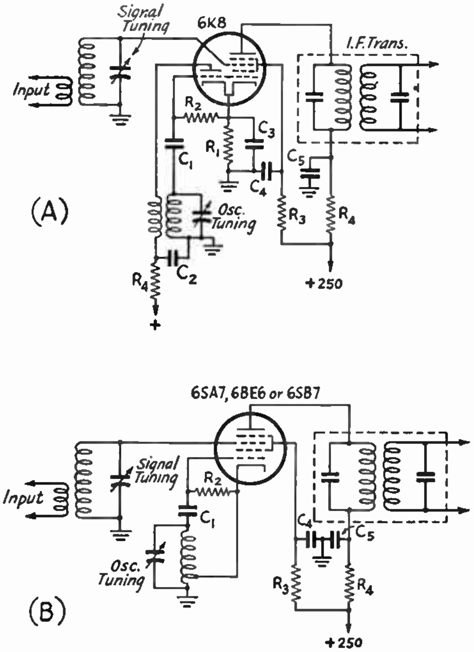


Fig. 5-12 — Typical circuits for triode-hexode (A) and pentagrid (B) converters. Values for R_1, R_2 and R_3 can be found in Table 5-I; others are given below.
C_1 — 47 μμfd.
C_3 — 0.01 μfd.
C_2, C_4, C_5 — 0.001 μfd.
R_4 — 1000 ohms.

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. 5-11 and 5-12.

The effectiveness of converter tubes of the type just described becomes less as the signal frequency is increased. Some oscillator voltage will be coupled to the signal grid through "space-charge" coupling, an effect that increases with frequency. If there is relatively little frequency difference between oscillator and signal, as for example a 14- or 28-Mc. signal and an i.f. of 455 kc., this voltage can become considerable because the selectivity of the signal circuit will be unable to reject it. The effect of this oscillator voltage on the grid is to make the output of the converter dependent upon the tuning of the signal circuit. If the signal grid is not returned directly to ground, but instead is returned through a resistor or part of an a.v.c. system, considerable bias can be developed which will also cut down the gain. For these reasons, and to reduce image response, the i.f. following the first converter of a receiver should be not less than 5 or 10 percent of the signal frequency, for best results.

Audio Converters

Converter circuits of the type shown in Fig. 5-12 can be used to advantage in the reception of c.w. and single-sideband suppressed-carrier signals, by introducing the local oscillator on the No. 1 grid, the signal on the No. 3 grid, and working the tube into an audio load. Its operation can be visualized as heterodyning the incoming signal into the audio range. The use of such circuits for audio conversion has been limited to selective i.f. amplifiers operating below 500 kc. and usually below 100 kc. An ordinary a.m. signal cannot be received on such a detector unless the tuning is adjusted to make the local oscillator zero-beat with the incoming carrier.

Since the beat oscillator modulates the electron stream completely, a large beat-oscillator component exists in the plate circuit. To prevent overload of the following audio amplifier stages, an adequate i.f. filter must be used in the output of the converter.

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

TABLE 5-I

Circuit and Operating Values for Converter Tubes

Plate voltage = 250 Screen voltage = 100, or through specified resistor from 250 volts

	SELF-EXCITED				SEPARATE EXCITATION			
Tube	*Cathode Resistor*	*Screen Resistor*	*Grid Leak*	*Grid Current*	*Cathode Resistor*	*Screen Resistor*	*Grid Leak*	*Grid Current*
6BA7[1]	0	12,000	22,000	0.35 ma.	68	15,000	22,000	0.35 ma.
6BE6[1]	0	22,000	22,000	0.5	150	22,000	22,000	0.5
6K8[2]	210	27,000	17,000	0.15-0.2	—	—	—	—
6SA7[2](7Q7[3])	0	18,000	22,000	0.5	150	18,000	22,000	0.5
6SB7Y[2]	0	15,000	22,000	0.35	68	15,000	22,000	0.35

[1] Miniature tube [2] Octal base, metal. [3] Lock-in base.

number of high-wattage resistors in the receiver and putting them in the separate power supply, and not mounting the oscillator coils and tuning condenser too close to a tube. Propping up the lid of a receiver will often reduce drift by lowering the terminal temperature of the unit.

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 power should be as low as is consistent with adequate putput. Low plate power 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.

Circuits

Several oscillator circuits are shown in Fig. 5-13. 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 a.c.-heated-cathode tubes are used. The circuit of Fig. 5-13C 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.

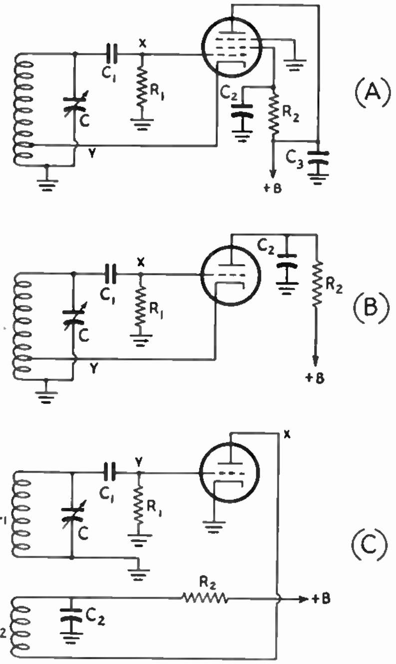


Fig. 5-13 — High-frequency oscillator circuits. A, pentode grounded-plate oscillator; B, triode grounded-plate oscillator; C, triode oscillator with tickler circuit. Coupling to the mixer may be taken from points X and Y. In A and B, coupling from *Y* will reduce pulling effects, but gives less voltage 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.

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 may cause the oscillator to "squeg" and generate 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.

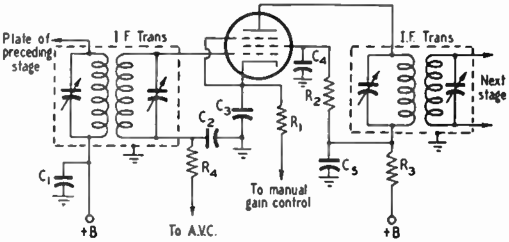


Fig. 5-14 — Typical intermediate-frequency amplifier circuit for a superheterodyne receiver. Representative values for components are as follows:

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.

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. Shifting the i.f. or better shielding are the solutions to such an interference problem.

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 at least be capable of amplifying equally all frequencies contained in a band extending from 5000 cycles above or 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 5 kc. wide, when the carrier is set at one edge. If the carrier is set in the center, a 10-kc. band would be required. 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, some of the sidebands are "cut." While sideband cutting 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 never serious with two-stage amplifiers at frequencies as low as 455 kc. A two-stage i.f. amplifier at 85 or 100 kc. will be sharp enough to cut some of the higher-frequency sidebands, if good transformers are used. However, the cutting is not at all serious, and the gain in selectivity is worthwhile if the receiver is used in the lower-frequency bands.

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 'phone reception.

A typical circuit arrangement is shown in Fig. 5-14. 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.

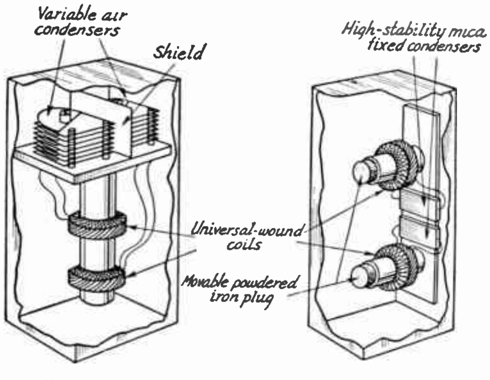
In Fig. 5-14, the gain of the stage is reduced by introducing a negative voltage to the lead marked "to a.v.c." or a positive voltage to R_1 at the point marked "to manual gain control." In either case, the voltage increases the bias on the tube and reduces the mutual conductance and hence the gain. When two or more stages are used, these voltages are generally obtained from common sources. The decoupling resistor, R_3, helps to isolate the amplifier from the power supply and thus prevents stray feed-back. C_2 and R_4 are part of the automatic volume-control circuit (described later); if no a.v.c. is used, the lower end of the i.f.-transformer secondary is simply connected to ground.

In a two-stage amplifier the screen grids of both stages may be fed from a common supply, either through a resistor (R_2) as shown, the screens being connected in parallel, or from a voltage divider across the plate supply. Separate screen voltage-dropping resistors are preferable for preventing undesired coupling between stages.

Typical values of cathode and screen resistors for common tubes are given in Table 5-II. The 6K7, 6SK7, 6BJ6 and 7H7 are recommended for i.f. work. The indicated screen resistors drop the plate voltage to the correct screen voltage, as R_2 in Fig. 5-14.

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
6BJ6	250	100	82	47,000
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



AIR TUNED PERMEABILITY TUNED

Fig. 5-15 — 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.

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 *Q*s and, hence, greater selectivity and gain per unit. In universal windings the coil is wound in layers with each turn traversing the length of the coil, back and forth, rather than being wound perpendicular to the axis as in ordinary single-layer coils. In a straight multilayer winding, a fairly large capacity can exist between layers. Universal winding, with its "criss-crossed" turns, tends to avoid building up such potential differences, and hence reduces distributed-capacity effects.

Variable tuning condensers are of the midget type, air-dielectric condensers being preferable because their capacity is practically unaffected by changes in temperature and humidity. Iron-core transformers may be tuned by varying the inductance (permeability tuning), in which case stability comparable to that of variable air-condenser tuning can be obtained by use of high-stability fixed mica condensers. Such stability is of great importance, since a circuit whose frequency "drifts" with time eventually will be tuned to a different frequency than the other circuits, thereby reducing the gain and

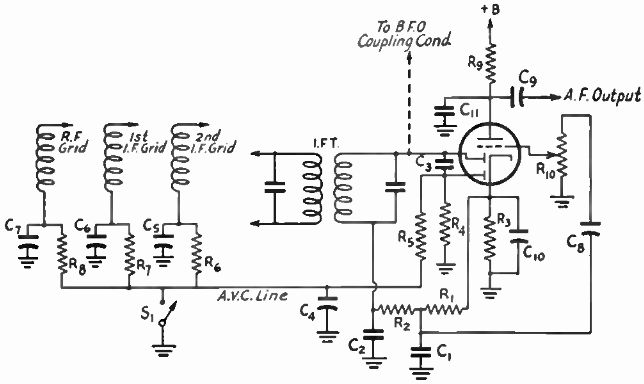


Fig. 5-16 — Automatic volume-control circuit using a dual-diode-triode as a combined a.v.c. rectifier, second detector and first a.f. amplifier.

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.

selectivity of the amplifier. Typical i.f.-transformer construction is shown in Fig. 5-15.

Besides the type of i.f. transformer shown in Fig. 5-15, 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:

	Bandwidth in Kilocycles		
Intermediate Frequency	*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. 5-13A 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. The b.f.o. power should be as low as is consistent with sufficient audio-frequency output on the strongest signals. However, if the beat-oscillator output is too low, strong signals will not give a proportionately strong audio signal. Contrary to some opinion, a weak b.f.o. is never an advantage.

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

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. The average rectified d.c. voltage, developed by the received signal across a resistance in a detector circuit, is used 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. 5-16. 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. 5-16. 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. 5-16 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 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. 5-16 will give a time constant that is satisfactory for average 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.

Amplified A.V.C.

The a.v.c. system shown in Fig. 5-16 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. 5-16, the

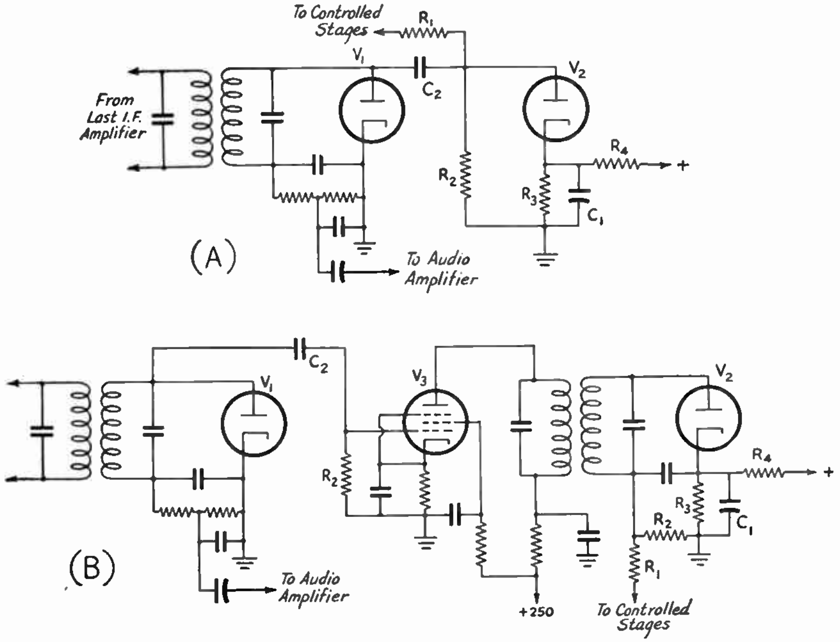


Fig. 5-17 — Delayed a.v.c. is shown at A, and amplified and delayed a.v.c. is shown in B. The circuit at B gives excellent a.v.c. action over a wide range, with no impairment of sensitivity for weak signals. For either circuit, typical values are:

C_1 — 0.001 μfd.
C_2 — 100 μμfd.
R_1, R_2 — 1.0 megohm.
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.

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. 5-17A. 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. 5-17B. A system like this, often called an "amplified a.v.c." system, gives superlative control action, since it maintains full receiver sensitivity for weak signals and substantially uniform audio output over a very wide range of signal strengths.

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).

The only known approach to reducing tube and circuit noise is through better "front-end" design and through more over-all selectivity.

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 desired signal 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.

Another approach is to "silence" (render inoperative) the receiver during the short duration time of any individual pulse. The listener will not hear the "hole" because of its short duration, and very effective noise reduction is obtained. Such devices are called "silencers" rather than "limiters."

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 during 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. 5-18 "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. 5-18A, 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, the circuit arrangement shown in Fig. 5-18B must be used.

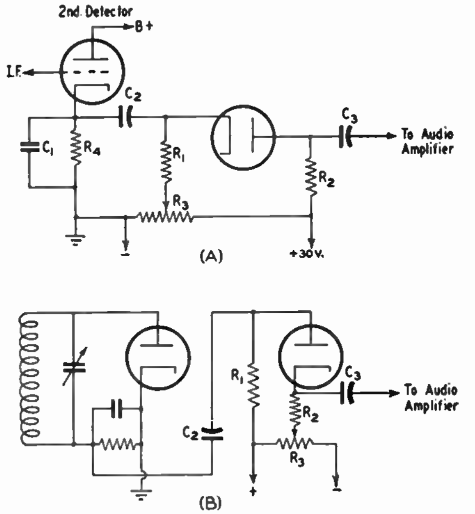


Fig. 5-18 — Series-valve noise-limiter circuits. A, as used with an infinite-impedance detector; B, with a diode detector. Typical values for components are as follows:

R_1 — 0.27 megohm.
R_2 — 47,000 ohms.
R_3 — 10,000 ohms.
R_4 — 20,000 to 50,000 ohms.
C_1 — 270 μμfd.
C_2, C_3 — 0.1 μfd.

All other diode-circuit constants in B are conventional.

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. 5-19. 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

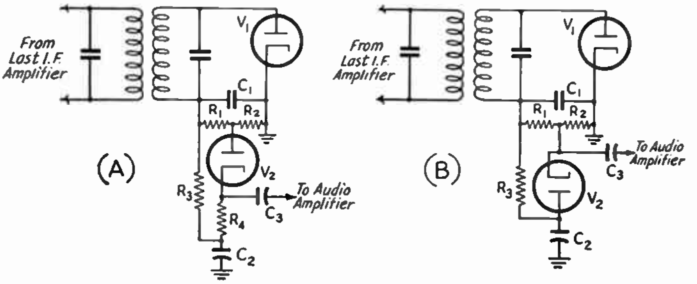


Fig. 5-19 — Self-adjusting series (A) and shunt (B) noise limiters. The functions of V_1 and V_2 can be combined in one tube like the 6H6 or 6AL5, or Type 1N31 crystals can be used.

C_1 — 100 μμfd.
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. 5-20, 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.

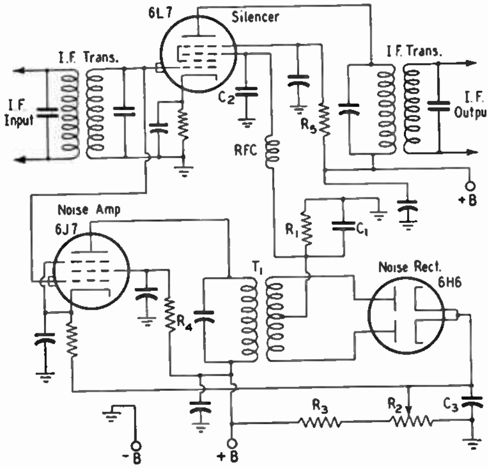


Fig. 5-20 — I.f. noise-silencing circuit. The plate supply should be 250 volts. Typical values for components are:

C_1 — 50-250 μμfd. (use smallest value possible without r.f. feed-back).
C_2 — 47 μμfd.
C_3 — 0.1 μfd.
R_1, R_4, R_5 — 0.1 meg.
R_2 — 5000-ohm variable.
R_3 — 22,000 ohms.
RFC — 20 mh.
T_1 — Special i.f. transformer for noise rectifier.

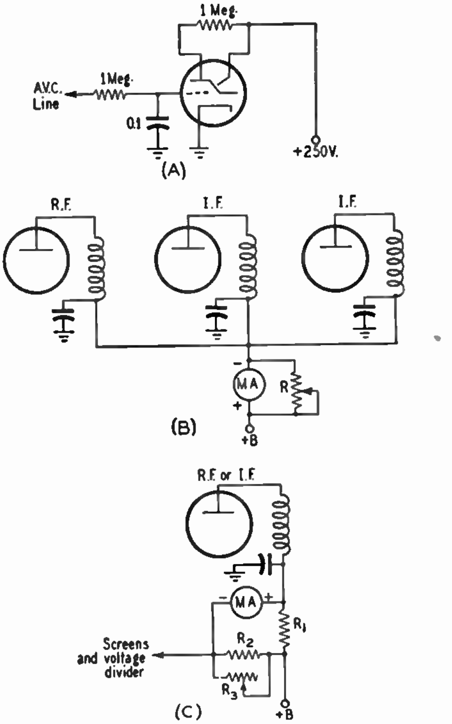


Fig. 5-21 — Tuning-indicator or S-meter circuits for superhet receivers. A, electron-ray indicator; B, plate-current meter for tubes on a.v.c.; C, bridge circuit for a.v.c.-controlled tube. In B, resistor *R* should have a maximum resistance several times that of the milliammeter. In C, representative 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. 5-21. 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 some 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

Probably the simplest means for 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. 5-22 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. 5-23. 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.

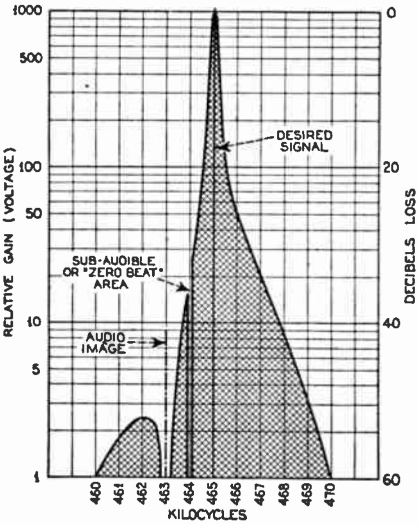


Fig. 5-22— Graphical representation of single-signal selectivity. The shaded area indicates the over-all bandwidth, or region in which response is obtainable.

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. 5-22.

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-

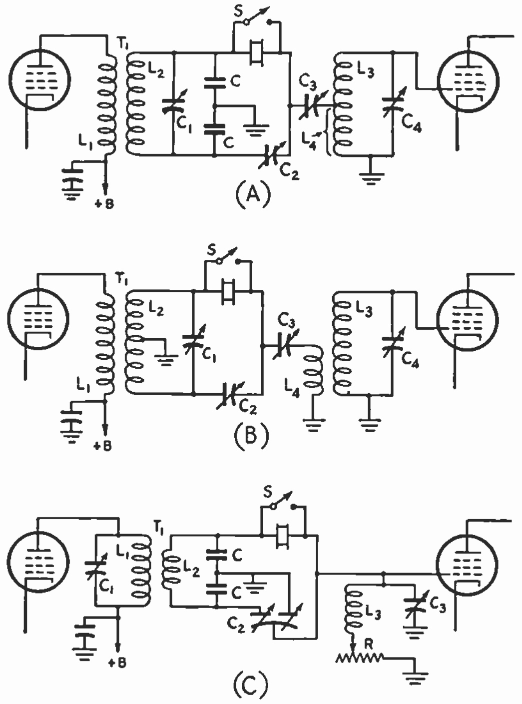


Fig. 5-23 — Crystal-filter circuits of three types. All give variable bandwidth, with C having the greatest range of selectivity. Suitable circuit values are as follows: Circuit A, T_1, special i.f. input transformer with high-inductance primary, L_1, closely coupled to tuned secondary, L_2; C_1, 50-μμfd. variable; C, each 100-μμfd. fixed (mica); C_2, 10- to 15-μμfd. (max.) variable; C_3, 50-μμfd. 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-μμfd. 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-μμfd. fixed (mica); C_2, opposed stator phasing condenser, approximately 8-μμfd. 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. 5-24.

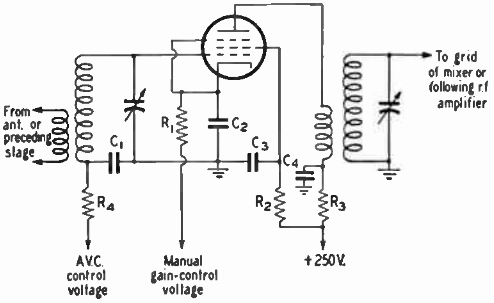


Fig. 5-24 — Typical radio-frequency amplifier circuit for a superheterodyne receiver. Representative values for components are as follows:

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.

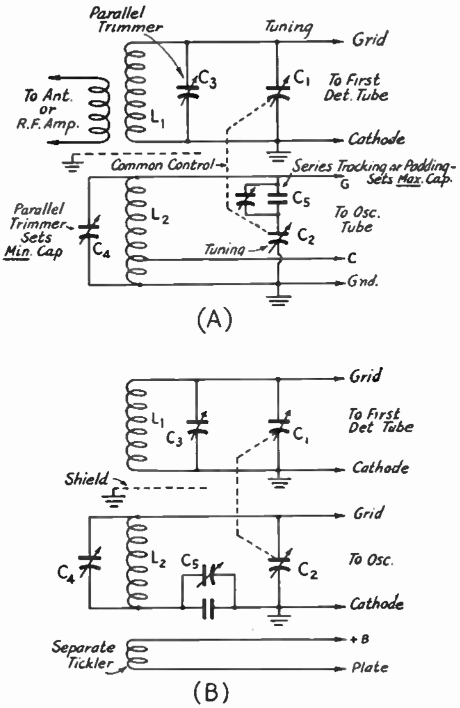


Fig. 5-25 — Converter-circuit tracking methods. Following are approximate circuit values for 450- to 465-kc. i.f.s, with tuning ranges of approximately 2.15-to-1 and C_2 having 140-$\mu\mu$fd. maximum, and the total minimum capacitance, including C_3 or C_4, being 30 to 35 $\mu\mu$fd.

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.	1.5 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. 5-25. 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. 5-25. The coils can be calculated with the *ARRL Lightning Calculator* and then 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. 5-8A, the tuning will be practically straight-line-frequency if C_2 (bandset) is 4 times or more the maximum capacity of C_1 (bandspread), as is usually the case for strictly amateur-band coverage. C_1 should be of the straight-line-capacity type (semi-circular plates).

Improving Receiver Sensitivity

Early in this chapter it was pointed out that the sensitivity (signal-to-noise ratio) of a receiver on the higher frequencies above 20 Mc. is dependent upon the bandwidth of the receiver and the noise contributed by the "front end" of the receiver. Neglecting the fact that the image rejection is poor, a receiver with no r.f. stage is generally satisfactory, from a sensitivity point, in the 3.5- and 7-Mc. bands. However, as the frequency is increased and the atmospheric noise becomes less, the advantage of a good "front end" becomes apparent. Hence at 14 Mc. and higher it is worth while to use at least one stage of r.f. amplification ahead of the first detector for best sensitivity as well as image rejection. The multigrid converter tubes have very poor noise figures, and even the best pentodes and triodes are three or four times noisier when used as mixers as they are when used as amplifiers.

If the purpose of an r.f. amplifier is to improve the receiver noise figure at 14 Mc. and higher, a high-G_m pentode or triode should be used. Among the pentodes, the best tubes are the 6AC7, 6AK5 and the 6SG7, in the order named. The 6AK5 takes the lead 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 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 coupled through a condenser to the grid of the last i.f. amplifier tube. 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 tuned 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 may 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. 5-26, 5-27, 5-28 and 5-29 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.

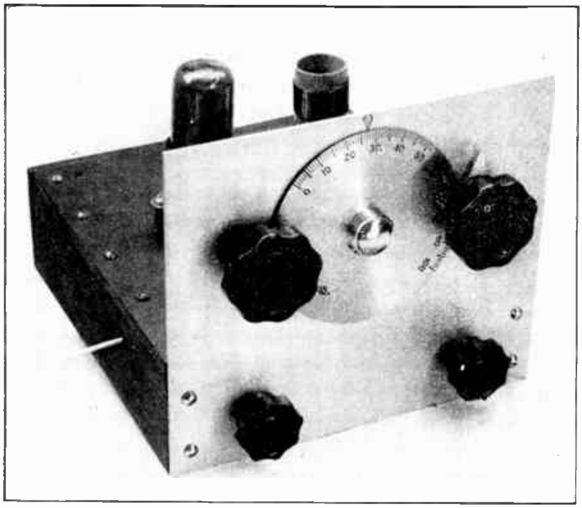


Fig. 5-26 — The simple one-tube regenerative receiver is built on a wood-and-Presdwood chassis, with an aluminum panel. The large left-hand knob drives the calibrated scale on the bandspread condenser. The large right-hand knob is for the band-set condenser.

From the circuit in Fig. 5-28, 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-

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Fig. 5-27 — 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.

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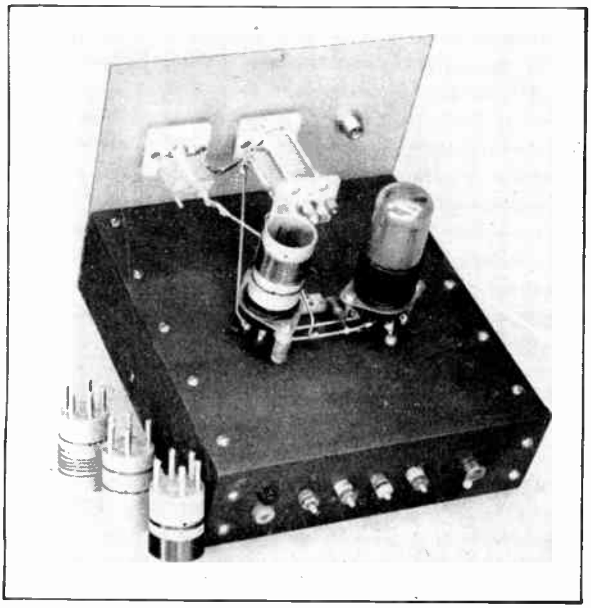

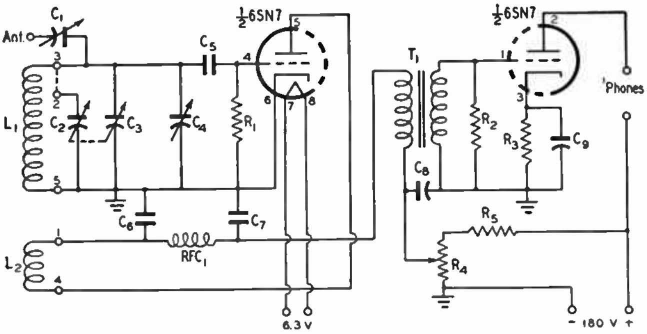

Fig. 5-28 — Wiring diagram of the one-tube regenerative receiver.

C_1 — Homemade adjustable condenser. See text.
C_2, C_3 — Reworked midget variable (Millen 21935). See text.
C_4 — 100-μμfd. midget variable (Millen 20100).
C_5 — 100-μμfd. mica.
C_6, C_7 — 470-μμfd. 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 μμfd. 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

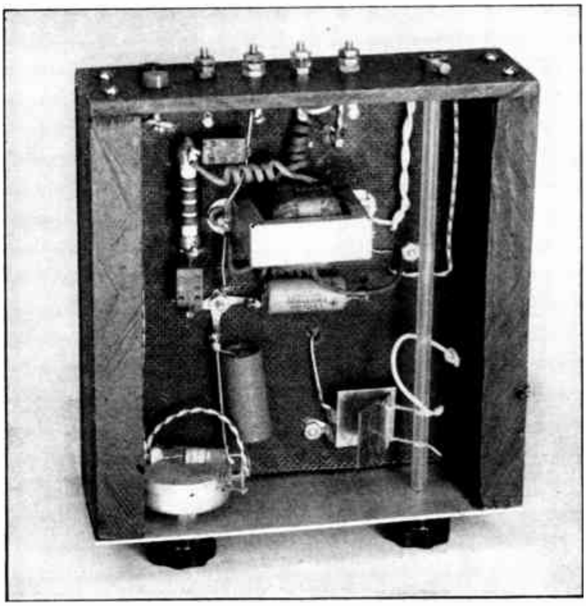

Fig. 5-29 — This view underneath the one-tube regenerative receiver shows the arrangement of parts and the construction of the variable antenna-coupling condenser.

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 ⅜ 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 ⅜ 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

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. 5-30 and 5-31, 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 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. 5-31 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. 5-31). Adding 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.

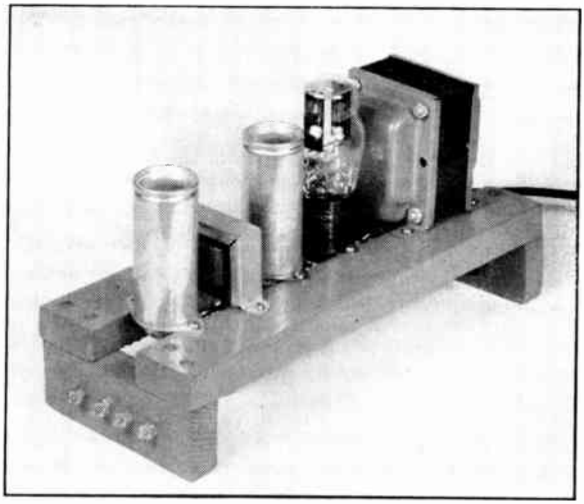

Fig. 5-30 — The power supply for the regenerative receiver is built on a simple wooden chassis.

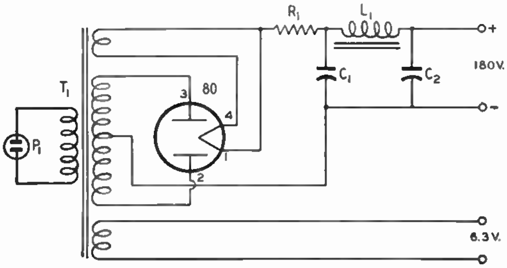

Fig. 5-31 — Circuit diagram of the power supply for the regenerative receiver.

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).

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. 5-32, 5-34 and 5-35. As can be seen from the circuit in Fig. 5-33, 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.

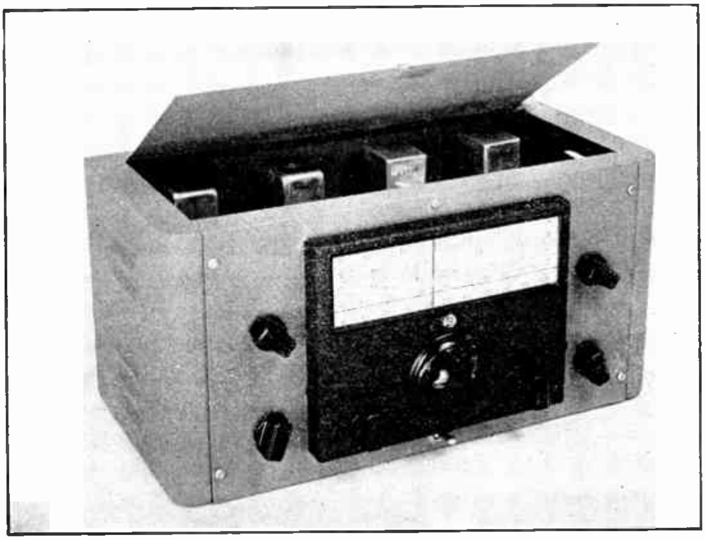

Fig. 5-32 — An amateur-band eight-tube receiver. The knobs on the left control audio volume (upper) and b.f.o. pitch, and the two on the right handle r.f. and i.f. gain (upper) and i.f. regeneration. The knob to the left of the large tuning knob is fastened to the *MAN.-A.V.C.-B.F.O* switch, and the 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.

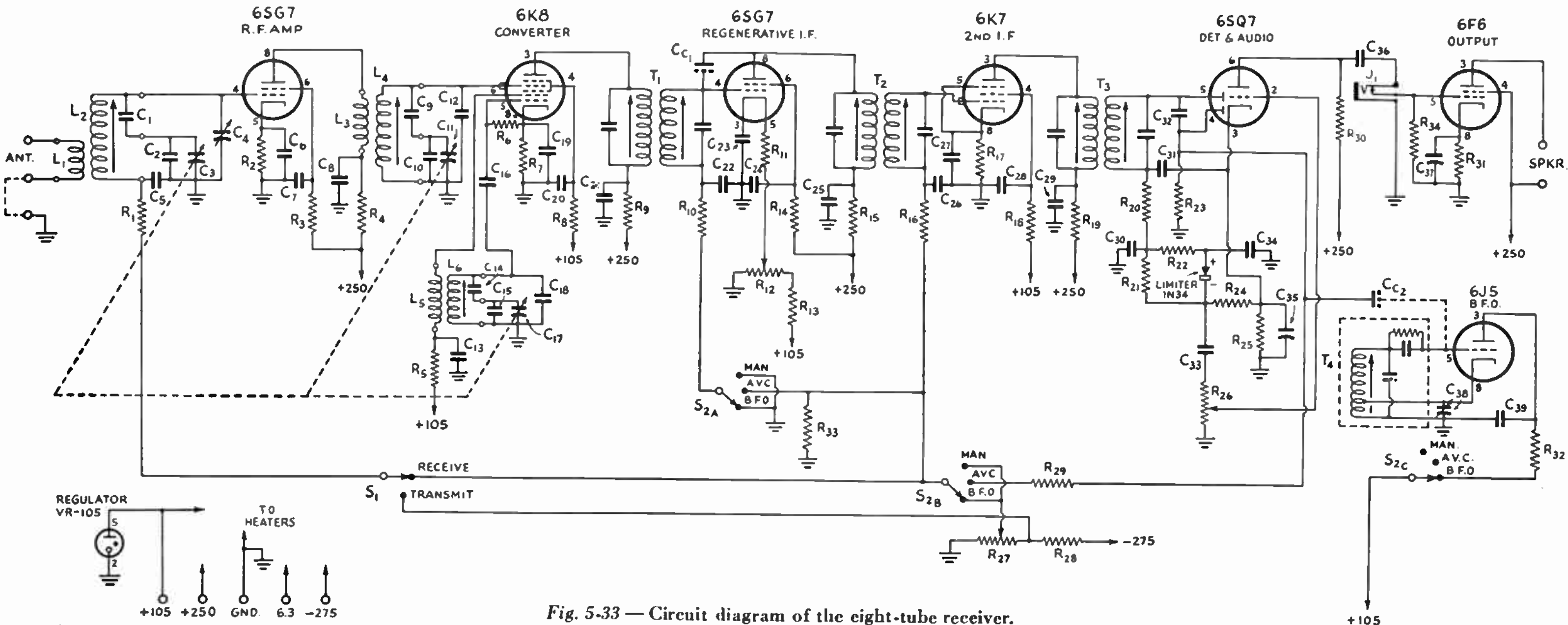

Fig. 5-33 — Circuit diagram of the eight-tube receiver.

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 64154).
T_4 — 456-kc. b.f.o. assembly, permeability-tuned (Millen 65456).

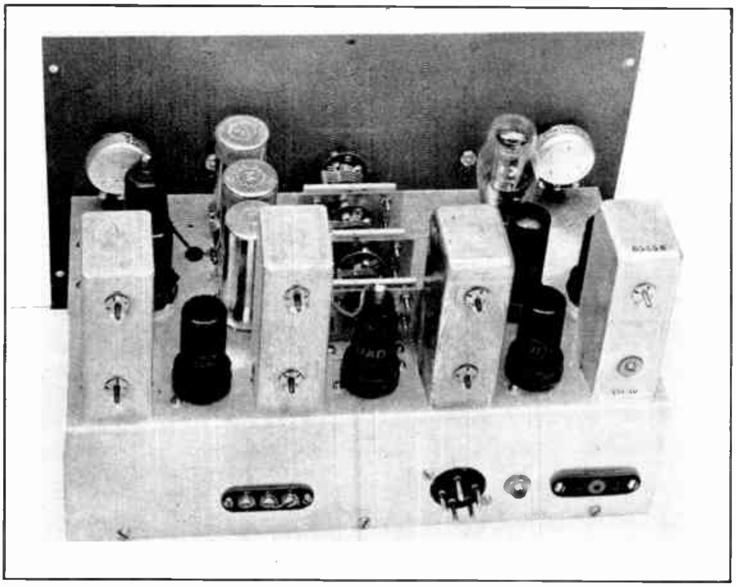

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Fig. 5-34 — This view of the eight-tube receiver chassis shows the mounting of the tuning condensers and the placement of most of the large components. The three shielded plug-in coil assemblies can be seen to the left of the tuning gang. The 6K8 converter 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.

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Construction

The receiver is built on an aluminum chassis mounted in a Par-Metal CA-202 cabinet, and a Millen 10035 dial is used for tuning. The chassis is made of 1/16-inch-thick stock, bent into a "U"-channel, and measures 13 inches wide and 7 1/4 inches deep on the top. It is 3 3/8 inches deep at the rear and 1/8 inch less at the front. The rear edge is reinforced with a piece of 3/8-inch square dural rod that is tapped for screws through the bottom of the cabinet, further to add to the strength of the structure when finally assembled. The various components that are common to the front lip of the chassis and the panel are used to tie the two together.

The shield panel used to mount the antenna-compensator condenser is also made of 1/16-inch aluminum with a 5/8-inch lip on the side for mounting. Part of the lip must be cut away to clear wires and mounting plates on some sockets, so it is advisable to put in the panel after most of the assembly and wiring have been completed. Flexible couplings and bakelite rod couple the condenser to the panel bushing.

The three tuning condensers are mounted on individual brackets of 1/16-inch aluminum. The brackets measure 2 1/2 inches wide and 1 9/16 high, with 1/2-inch lips. A cover of thin aluminum — not shown in the photographs — slides over the condenser assembly to dress up the top view a bit. The dust cover is not necessary for satisfactory operation of the receiver.

Ceramic sockets are used for the plug-in coils and for the r.f. amplifier, converter and b.f.o. tubes. Mica condensers were used throughout the receiver for by-passing wherever feasible, because they lend themselves well to compact construction. Ceramic condensers are now available that could be used in the i.f. amplifier at considerable saving in cost and room.

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.

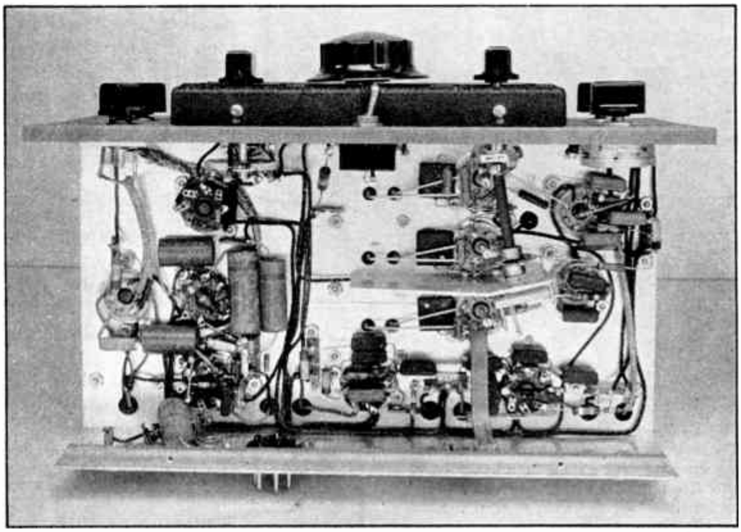

Fig. 5-35 — The by-pass condensers used throughout the r.f. and i.f. stages are grouped around the sockets of their respective tubes. Tie-points are used wherever necessary to support small resistors and condensers. The antenna trimmer condenser is mounted on a bracket which also serves as shielding 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.

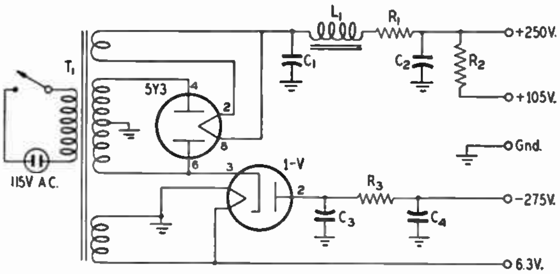

Fig. 5-36 — Wiring diagram of power supply for the eight-tube receiver.
C_1, C_2 — 16-μfd. 450-volt electrolytic.
C_3, C_4 — 8-μfd. 450-volt electrolytic.
R_1 — 500 ohms, 10 watts, wire-wound.
R_2 — 5000 ohms, 10 watts, wire-wound.
R_3 — 0.1 megohm, 1 watt, composition.
L_1 — 30-henry 110-ma. filter choke (Stancor C-1001).
T_1 — 350–0–350 volts, 90 ma.; 5 volts at 3 amp., 6.3 volts at 3.5 amp.

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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.

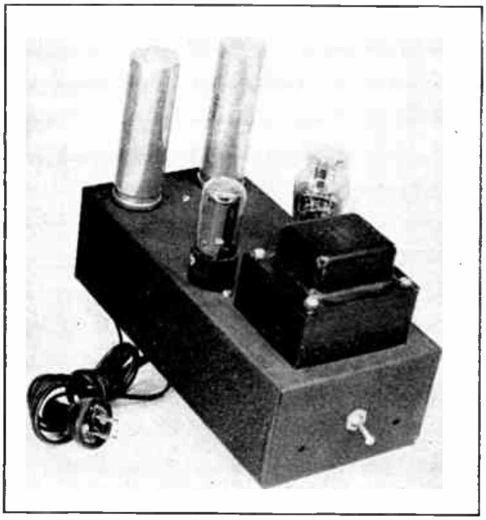

Fig. 5-37 — Power supply for the eight-tube receiver. Two rectifiers are required because a separate supply is incorporated for gain-control purposes. The filter choke and the negative-supply filter condensers are mounted under the chassis. At the rear of the chassis is the socket for the power cable.

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 "MAX." 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. 5-36 and 5-37. An idea of the parts arrangement can be obtained from Fig. 5-37, 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_{35} 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 Clipper/Filter For C.W. or 'Phone

The clipper/filter 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 or c.w. signals, it will keep the strength of c.w. signals at a constant level, and it will add selectivity to your receiver for c.w. reception. 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.

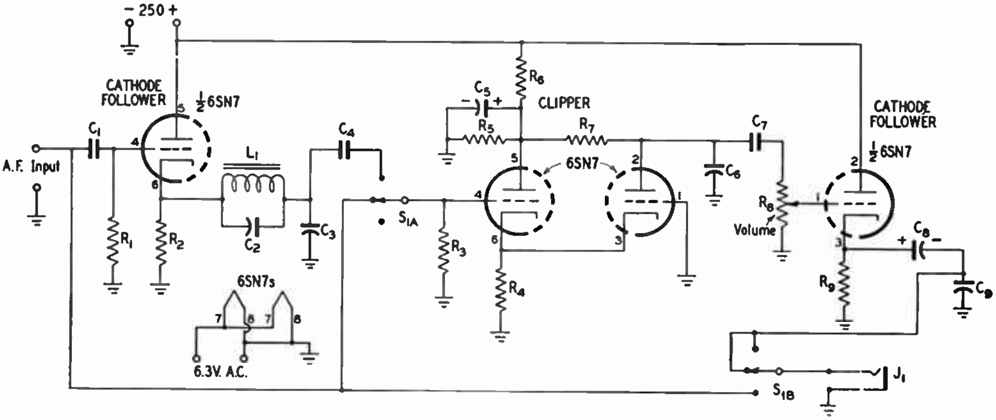

Fig. 5-38 — Circuit diagram of the audio clipper unit. Power requirements are 16 ma. at 250 v. d.c., 1.2 amp. at 6.3 v. a.c.

C_1, C_4, C_7 — 470-$\mu\mu$fd. mica.
C_2 — 0.04-μfd. paper.
C_3 — 0.1-μfd. paper.
C_5 — 8-μfd. 450-volt electrolytic.
C_6 — 0.003-μfd. paper.
C_8 — 10-μfd. 25-volt electrolytic.
C_9 — 0.25-μfd. paper.
R_1, R_3 — 1 megohm, ½ watt.
R_2, R_9 — 1500 ohms, ½ watt.
R_4 — 10,000 ohms, ½ watt.
R_5 — 22,000 ohms, ½ watt.
R_6 — 47,000 ohms, 1 watt.
R_7 — 33,000 ohms, ½ watt.
R_8 — 1-megohm volume control.
L_1 — 250-mh. choke (Millen 34400-250).
J_1 — 'Phone jack, single circuit.
S_1 — 2-circuit 3-position switch.

The circuit is shown in Fig. 5-38. The constants are not too critical, and have been adjusted for operation at the signal levels ordinarily available from the headphone jack on a receiver. The clipper output circuit is heavily by-passed by C_6 to reduce the amplitude of the harmonics generated in the clipping process, and additional by-passing by C_9, across the headset, is used for the same purpose. Cathode-follower input and output circuits allow the unit to be used with any receiver output and any headphones, and they also

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Fig. 5-39 — The audio clipper unit includes input and output amplifiers of the cathode-follower type, a dual-triode clipper circuit, and a selective audio system. It is built in a small utility box, with a cable for power-supply connections and a cord and plug to pick up audio from the receiver's headphone jack.

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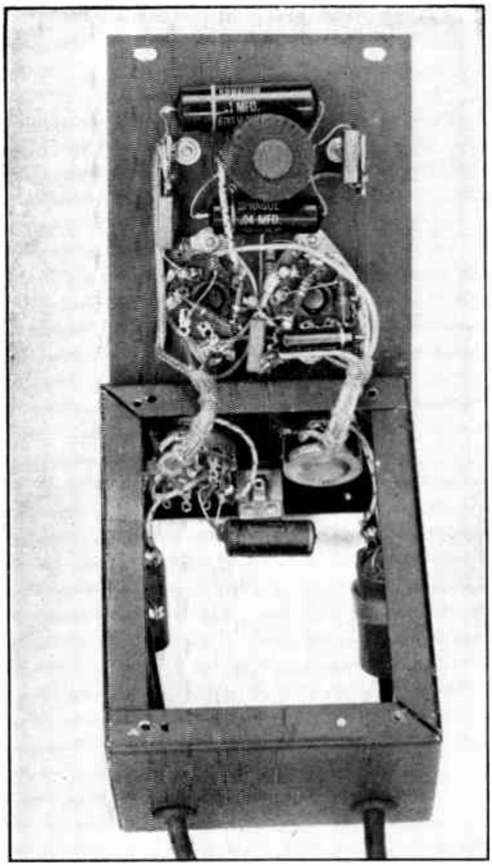

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Fig. 5-40 — Inside view of the clipper unit. The gain control, switch, headphone jack, and the larger fixed condensers are mounted on the walls of the box. The two tubes and the selective audio circuit are mounted on the removable panel. The selective circuit, consisting of the choke coil and two tubular condensers, occupies the upper half of the panel in this view. The socket at the left is for the input and output amplifiers: the right-hand socket is for the double-triode clipper.

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contribute to the effectiveness of the audio filter, $L_1C_2C_3$. A three-position switch, S_1, is provided so that the unit can be cut out entirely, used with straight limiting and no selectivity, or with both selectivity and limiting. The "off" position is useful principally to convince the skeptical, and the limiting without selectivity is useful for impulse noise, when encountered. High selectivity and good noise suppression do not go hand in hand.

The unit, shown in Figs. 5-39 and 5-40, is built on one panel and the sides of a 3 by 4 by 5 utility box. The parts on the panel and the box proper are connected through cabled leads made long enough so the panel can be swung out as shown. Any type of construction can be used, since there is nothing critical in the layout. One precaution to observe is to use a shielded lead between the "hot" input terminal and the switch, to prevent possible stray coupling between the input and later high-impedance circuits because of the cabled leads.

The selective audio circuit chosen gives a type of frequency-response curve that is quite useful. The peak at 800 cycles is broad enough to avoid tuning difficulties, even when used in conjunction with the crystal filter in the receiver. Nevertheless, the response drops off rapidly enough, particularly on the high-frequency side, to make a marked difference in respect to the "capturing" of the limiter by strong off-resonance signals. There is a "notch" at 1700 cycles.

There is a wide latitude in choice of inductances for L_1. The Millen coil listed under Fig. 5-38 was the best of available low-priced units tried, in terms of sharpness of the response curve and the depth of the rejection notch. Some of the small filter chokes such as the Stancor C-1515 and Thordarson T20C53 also work reasonably well. The former will resonate at approximately the same frequencies as given above with 330 $\mu\mu$fd. at C_2 and 470 $\mu\mu$fd. at C_3; the latter choke requires 0.001 μfd. at C_2 and 0.002 μfd. at C_3. With any coil the values of capacitance required to place the peak and notch at frequencies that best fit one's taste in beat notes can easily and quickly be determined by simple cut-and-try. Other types of selective audio circuits can, of course, also be substituted.

In use, the receiver's gain controls should be set so that only the stronger signals are clipped; too-deep clipping will make the receiver sound as though practically every signal overloads it. Once the proper settings for clipping level are determined, the actual audio volume is adjusted by the gain control on the unit. A little juggling back and forth between the receiver controls and the output control in the clipper unit will eventually result in the receiver's sounding very much like it does without the clipper present. The difference is that the signals and noise, including one's own transmitter signal, don't rise above the level set as a ceiling.

The "Selectoject"

The Selectoject is a receiver adjunct that can be used as a sharp amplifier or as a single-frequency rejection filter. The frequency of operation may be set to any point in the audio range by turning a single knob. The degree of selectivity (or depth of the null) is continuously adjustable and is independent of tuning. In 'phone work, the rejection notch can be used to reduce or eliminate a heterodyne. In c.w. reception, interfering signals may be rejected or, alternatively, the desired signal may be picked out and amplified. The Selectoject may also be operated as a low-distortion variable-frequency audio oscillator suitable for amplifier frequency-response measurements, modulation tests, and the like, by advancing the "selectivity" control far enough in the selective-amplifier condition. The Selectoject is connected in a receiver between the detector and the first audio stage. Its power requirements are 4 ma. at 150 volts and 6.3 volts at 0.6 amperes. For proper operation, the 150 volts should be obtained from across a VR-150 or from a supply with an output capacity of at least 20 μfd.

The wiring diagram of the Selectoject is shown in Fig. 5-41. Resistors R_2 and R_3, and R_4 and R_5, can be within 10 percent of the nominal value but they should be as close to each other as possible. An ohmmeter is quite satisfactory for doing the matching. One-watt resistors are used because the larger ratings are usually more stable over a long period of time.

If the station receiver has an "accessory socket" on it, the cable of the Selectoject can be made up to match the connections to the socket, and the numbers will not necessarily match those shown in Fig. 5-41. The lead between the second detector and the receiver gain control should be broken and run in shielded leads to the two pins of the socket corresponding to those on the plug marked "A.F. Input" and "A.F. Output." If the receiver has a VR-150 included in it for voltage stabilization there will be no problem in getting the plate voltage — otherwise a suitable voltage divider should be incorporated in the receiver, with a 20- to 40-μfd. electrolytic condenser connected from the +150-volt tap to ground.

In operation, overload of the receiver or the Selectoject should be avoided, or all of the possible selectivity may not be realized.

The Selectoject is useful as a means for obtaining much of the performance of a crystal filter from a receiver lacking a filter, and it is just as useful with a crystal filter in that it provides additional flexibility and performance. It is the answer to the 'phone man's prayer: "If I only had a second crystal filter!"

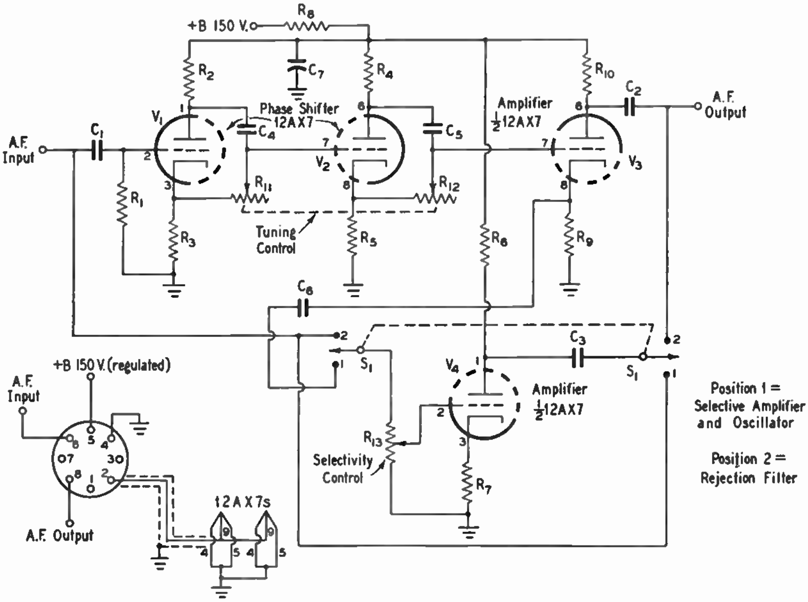

Fig. 5-41 — Complete schematic of Selectoject using 12AX7 tubes.

C_1 — 0.01-μfd. mica, 400 volts.
C_2, C_3 — 0.1-μfd. paper, 200 volts.
C_4, C_5 — 0.002-μfd. paper, 400 volts.
C_6 — 0.05-μfd. paper, 400 volts.
C_7 — 16-μfd. 150-volt electrolytic.
R_1 — 1 megohm, ½ watt.
R_2, R_3 — 2000 ohms, 1 watt, matched as closely as possible (see text).
R_4, R_5 — 4000 ohms, 1 watt, matched as closely as possible (see text).
R_6 — 20,000 ohms, ½ watt.
R_7 — 2000 ohms, ½ watt.
R_8 — 10,000 ohms, 1 watt.
R_9 — 6000 ohms, ½ watt.
R_{10} — 20,000 ohms, ½ watt.
R_{11}, R_{12} — Ganged 0.5-megohm ½-watt potentiometers, standard audio taper (tuning control).
R_{13} — 0.5-megohm ½-watt potentiometer (selectivity control).
S_1 — D.p.d.t. toggle.

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. 5-42 and 5-44 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. 5-43, 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. 5-42, 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. 5-44) 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.

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)

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

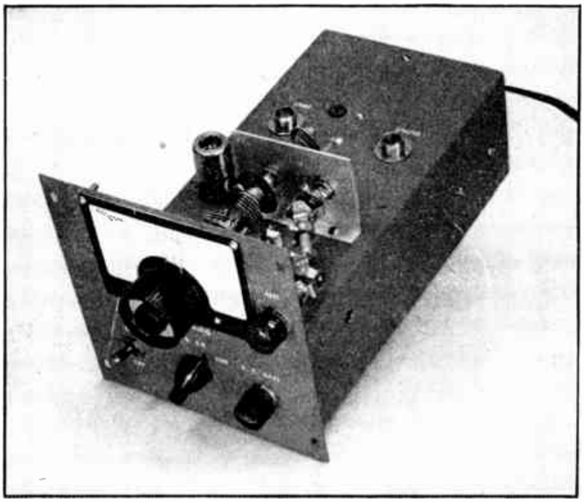

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Fig. 5-42 — A bandswitching preselector for 14 and 28 Mc. A single 6AK5 amplifier is used, and the power supply is included in the unit. The antenna-trimming condenser is mounted on the small aluminum partition.

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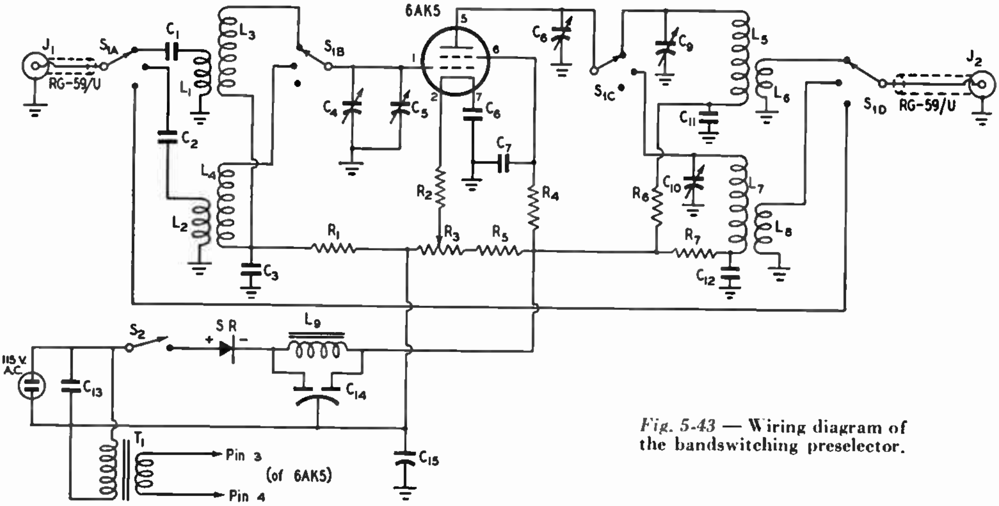

Fig. 5-43 — Wiring diagram of the bandswitching preselector.

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 — 4700 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-101).
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.

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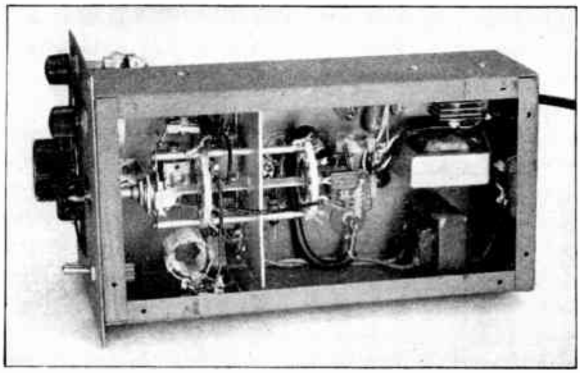

Fig. 5-44 — A view underneath the chassis of the bandswitching preselector, showing the shield partition between switch sections and the selenium rectifier and associated filter.

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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.

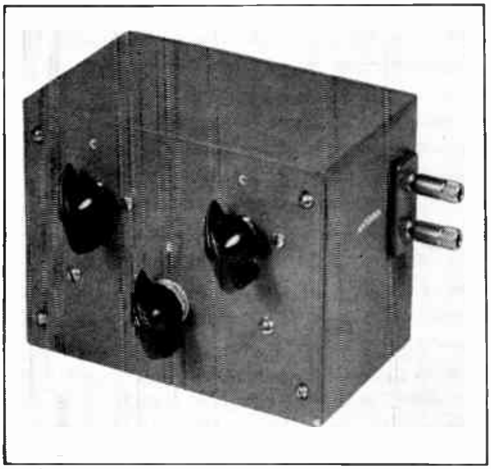

Fig. 5-45 — A compact coupling network for matching a balanced line to the receiver on 14 and 28 Mc.

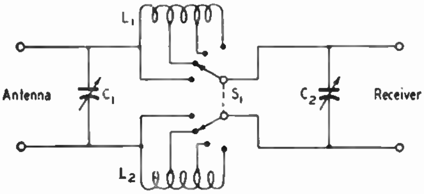

Fig. 5-46 — Circuit diagram of the coupling unit.

C_1, C_2 — 100-μμfd. midget variable (Millen 22100).

L_1, L_2 — 30 turns No. 24 d.c.c. close-wound on ⅝-inch diameter polystyrene form, tapped at 2½, 6½ and 14½ turns.

S_1 — 2-circuit 5-position single-section ceramic wafer switch (Mallory 173C).

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. 5-46. 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 PRE-3 polystyrene forms, are fastened to the panel with brass screws, and the coils should be wound on the forms as far away as possible from the mounting end. If this still leaves the coil ends within ½ inch of the panel, the forms should be spaced away from the panel by National XP-6 buttons. The switch should be wired so that the switching sequence puts in, in each coil, 0 turns, 2½ turns, 6½ turns, 14½ turns and 30 turns. All of the wiring, with the exception of the leads to the input and output terminals, can be done with the panel removed from the box.

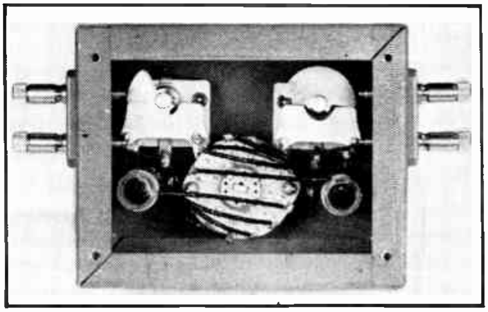

Fig. 5-47 — Rear view of the antenna-coupling unit. The two coils can be seen directly below the two tuning condensers.

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 Tuned Lines

The pi-section coupler shown in Figs. 5-45, 5-46 and 5-47 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. 5-48, 5-49 and 5-50, 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.

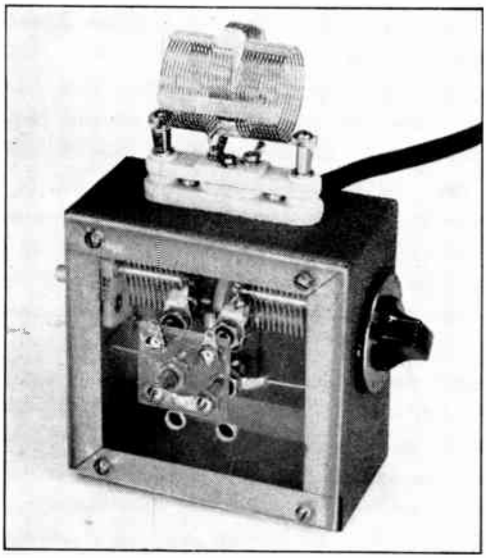

Fig. 5-48 — A small tuned coupler for matching the receiver to a tuned line. The unit is made either series- or parallel-tuned by the position of the antenna connection block.

As can be seen in the wiring diagram, Fig. 5-49, 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.

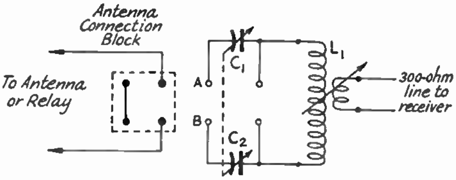

Fig. 5-49 — Circuit of the tuned antenna coupler.
C_1, C_2 — 100-μμfd. midget variable (Millen 22100).
L_1 — Coil to tune to band in use, with swinging link (National AR-16).

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.

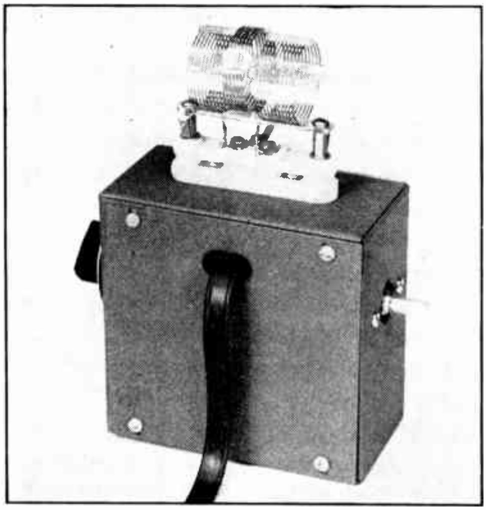

Fig. 5-50 — Another view of the tuned antenna coupler.

A One-Tube Converter For 10 and 11 Meters

The 10- and 11-meter converter shown in Figs. 5-51 and 5-53 is a simple unit that can be built in a few hours, for a cost of less than fourteen dollars. The converter uses a fixed-tune i.f. and tunable input and oscillator circuits, in preference to a fixed-frequency oscillator and a tunable output circuit. With a one-tube converter of the latter type, it is almost impossible to avoid picking up at least a few signals in the tuning range of the receiver. Using a tunable oscillator and a fixed-frequency output circuit permits one to select an i.f. free from interference. The plate-current demand is only 5 ma., and it is usually possible to operate the converter from the receiver power supply.

As can be seen in Fig. 5-52, the Hartley circuit is used in the oscillator portion of the 6BA7 pentagrid converter. A padding condenser, C_2, is switched in through S_1 to change the range for 11-meter operation. Condenser C_4 is used for tuning, and the input circuit is tuned to either range with C_1. The screen grid of the 6BA7 is operated at about 65 volts, since higher voltages will increase the total tube current without any marked improvement in performance. However, since the available supply voltage will vary with different receivers, the value of the screen dropping resistor, R_2, cannot be specified, and it must be calculated, as described later.

There is a good reason for not using an antenna switch for straight-through operation of the converter. With practically any available switch it is very difficult to prevent capacity coupling between the input and output circuits of the converter. Any such capacity coupling increases the problem of eliminating interference at the i.f. By equipping the converter and the receiver with identical input terminals and using similar plugs on both the antenna feed line and the converter output cable, antenna changeover is no problem. The metal partition separating L_2 and L_3, shown in Fig. 5-52, reduces the effect of oscillator harmonics beating with high-frequency (f.m.) broadcast stations.

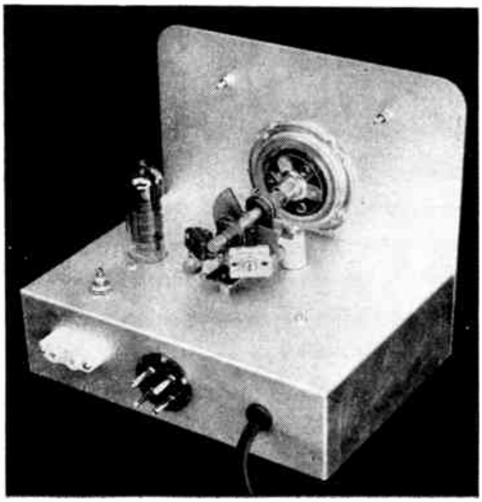

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Fig. 5-51 — A one-tube converter for extending the tuning range of a receiver to 10 and 11 meters. The crystal socket on the back of the chassis receives the antenna plug (Millen 37412).

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Construction

The converter is built on a 5 by 7 by 2-inch aluminum chassis, and a 6 by 7-inch panel is held in place by the components mounted on the front wall of the chassis. The main tuning dial is a National type MCN.

It can be seen in Fig. 5-51 that the oscillator tuning condenser, C_4, is mounted on ¼-inch metal pillars. A National Type GS-10 stand-off insulator is located at the front-right-hand side of C_4, and a soldering lug at the top end of this insulator is soldered to the stator terminal lug of the condenser. This added support for the tuning condenser improves oscillator stability, by preventing rocking of C_4 as the control shaft is turned. A feed-through bushing at the other front terminal of the condenser is used to support and insulate the lead passing through the chassis to the coil below. The padder condensers for the oscillator circuit, C_3 and C_5, are mounted on the rear terminal lugs of the tuning condenser.

The grid coil, L_2, is mounted on the terminal lugs of the input tuning condenser, C_1. The antenna coil, L_1, should be wound around L_2 before the larger coil is soldered in place. The tube socket, to the rear of C_1L_2, is mounted with pins No. 1 and 7 facing toward the rear of the chassis. The aluminum shield between the input and the oscillator coils has a ⅜-inch lip bent over along one edge, for fastening to the chassis. The shield is slotted to clear the cathode-tap lead.

The screen and decoupling resistors, R_2 and R_3 — respectively, are supported at the power-supply ends by a tie-point strip which is held in place by the same screw that anchors the soldering lug for L_3. 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

$$\text{R (ohms)} = \frac{\text{supply voltage} - 65}{0.0046}$$

Example: Supply voltage 260; the resistor value is $\frac{260 - 65}{0.0046} = 42{,}391$ ohms. Anything within 20% of this figure would be satisfactory.

The coaxial output cable is terminated at the chassis end at a tie-point strip located at the left end of the chassis.

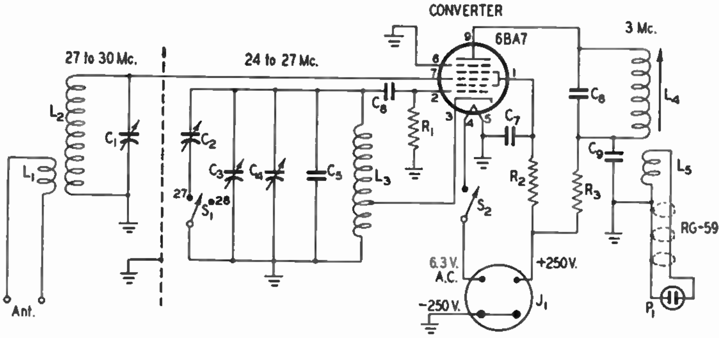

Fig. 5-52 — Circuit diagram of the low-cost 10- and 11-meter converter.

C_1 — 15-μμfd. variable (Millen 20015).
C_2, C_3 — 3-30-μμfd. mica trimmer.
C_4 — 25-μμfd. variable (Millen 19050 with 2 stator and 2 rotor plates removed).
C_5 — 68-μμfd. silver mica.
C_6 — 47-μμfd. ceramic.
C_7, C_9 — 0.01-μfd. disc ceramic.
C_8 — 82-μμfd. mica.
R_1 — 22,000 ohms, ½ watt.
R_2 — Screen resistor: see text.
R_3 — 1000 ohms, ½ watt.
L_1 — 3 turns No. 24 d.s.c., space wound around L_2.
L_2 — 13 turns No. 20 tinned, ⅝-inch diam., 1³⁄₁₆-inch long (B & W 3007).
L_3 — 6 turns No. 18 tinned, ½-inch diam., ¾-inch long, cathode tap 1¾ turns from ground end (B & W 3002).
L_4 — Slug-tuned plate coil (CTC LS3 — 5 MC.).
L_5 — 10 turns No. 24 d.s.c. scramble wound at cold end of L_4.
J_1 — Panel-mounting male socket (Amphenol 86-CP4).
P_1 — 300-ohm twin-lead plug (Millen 37412).
S_1, S_2 — S.p.s.t. toggle switch.

It is important that the link from the converter to the receiver be well shielded, to avoid picking up any signals directly in the receiver. A length of RG-58/U or RG-59/U can be used and, if necessary, a small shield should be mounted over the antenna binding post of the receiver. However, it is usually possible to set the receiver somewhere near 3 Mc. that will be free from even the weakest straight-through interference.

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.

Testing

Power for the converter can be obtained from a separate supply, but it is usually more convenient to "steal" the power from the receiver. The converter requires 6.3 volts at 0.3 amperes for the heater and 200 to 250 volts d.c. at 5 to 6 ma. for the plate and screen.

After the power supply has been connected, it is advisable to check the screen and plate voltages with a voltmeter. It may be necessary to change the value of R_2 if the screen voltage isn't in the recommended range of 60 to 70.

If your transmitter uses VFO, set the VFO to have a harmonic fall at 28 Mc., and tune the receiver to 3 Mc. If you have crystal control, turn on the oscillator and set the receiver to the crystal's 28-Mc. harmonic minus 25 Mc. If, for example, your crystal has a harmonic at 28,650 kc., set the receiver to 3650 kc. Set the tuning condenser, C_4, to where you want the test frequency (transmitter-oscillator harmonic) to appear on the dial, and tune it in by adjusting C_3. If the signal is too loud, remove any test antenna from the converter. With a reasonable signal, check the tuning of the input circuit, C_1L_2, and adjust L_4 for maximum signal in the receiver.

Once the converter has been set up on known frequencies within the 10- and 11-meter bands, C_2 and C_3 are left fixed and the tuning is done with C_4. The bandspread will be approximately 80 dial divisions on 10 and 20 or so on 11 meters. C_1 need not be touched over a tuning range of about 200 kc., and so should be used at intervals if the entire band is being combed.

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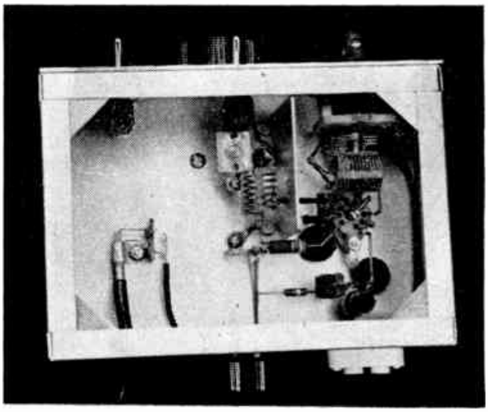

Fig. 5-53 — A bottom view of the one-tube converter. The toggle switches are for band-changing and opening the heater circuit.

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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. 5-54 and 5-56. 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. 5-55. 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-

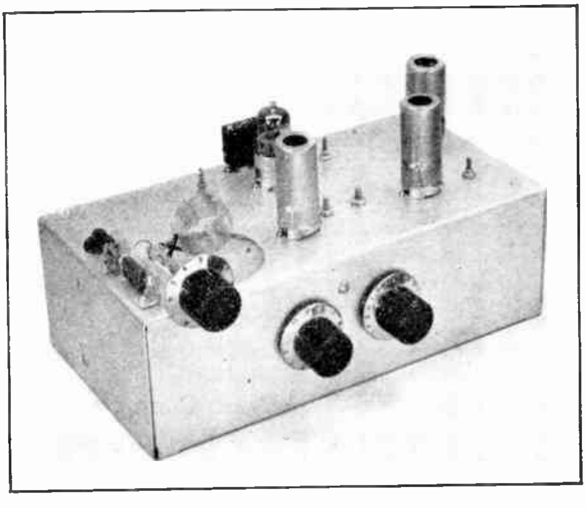

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Fig. 5-54 — **A 28-Mc. crystal-controlled converter. The adjustable antenna coupling can be seen at the left front. The tube shields, from left to right, cover the triode-connected 6AK5 r.f. amplifier, the 6AK5 mixer and the 6C4 cathode follower. The unshielded tube is the 6J6 oscillator-multiplier.**

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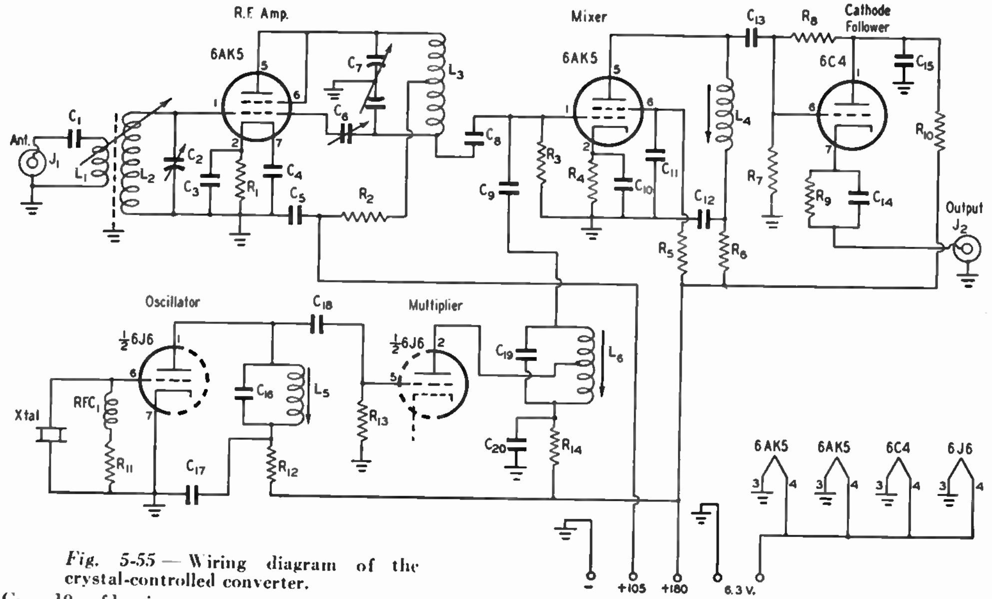

Fig. 5-55 — Wiring diagram of the crystal-controlled converter.

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. 5-54.) 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. 5-56 and 5-57 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. 5-57. 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.

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}	75μμfd.	75μμfd.	33μμfd.
C_{19}	0	22μμfd.	22μμfd.
Xtal	6000 kc. (triples)	5875 kc. (triples)	12,250 kc. (doubles)

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. Then back off on L_5 a little, because there is no need to run the crystal at maximum.

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. If your receiver has an antenna trimmer, peak it too. Then plug in the 6AK5 r.f. amplifier and, after the tube has warmed up, rock C_2 and C_7. Through the hole in the bottom plate, use an alignment tool to adjust C_6 a little at a time, until you lose any unpleasant sounds with all settings of C_2 and C_7, and the r.f. stage is neutralized. Connect the antenna, and peak C_2 and C_7 on a signal. 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

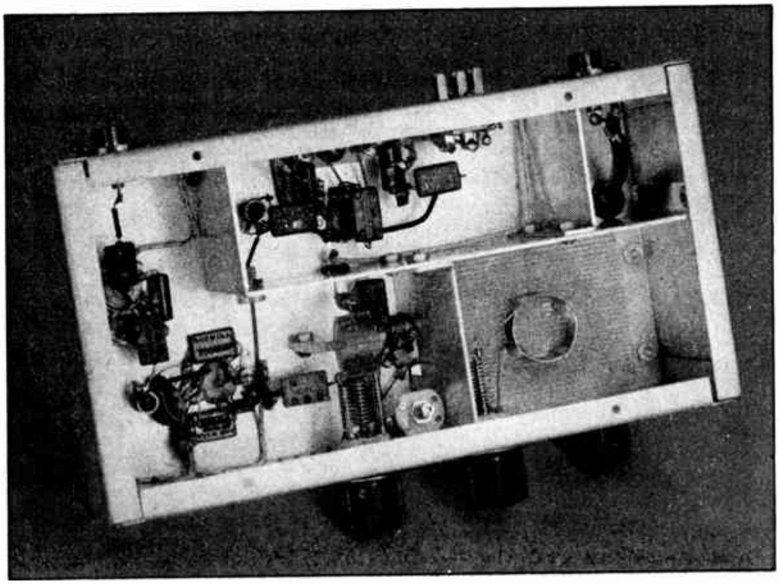

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Fig. 5-56 — This view of the underside of the converter with the bottom cover removed shows the Faraday shield at the lower right, the shield straddling the r.f. amplifier socket (lower center) and the shielded oscillator section (top center). The neutralizing condenser for the r.f. stage is adjusted through a hole in the bottom cover.

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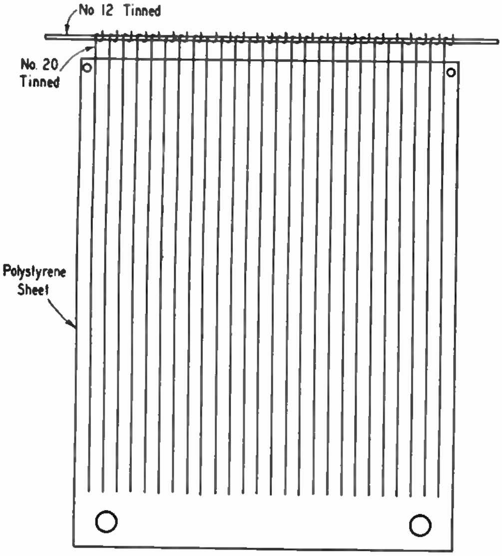

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Fig. 5-57 — Constructional details of the Faraday shield, before soldering the ends of the No. 20 wires to the No. 12 wire bus.

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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. 5-58, 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.

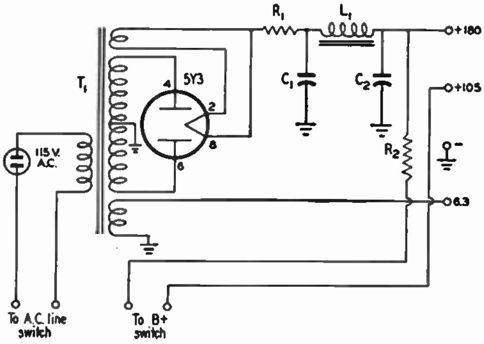

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Fig. 5-58 — A power supply for the crystal-controlled converter.

C1, C2 — 8-μfd. 450-volt electrolytic.
R1 — 1500 ohms, 10 watts.
R2 — 10,000 ohms, 10 watts.
L1 — 16-hy. 50-ma. choke (Stancor C-1003).
T1 — 240-0-240 at 40 ma., 5 and 6.3 v. (Stancor P-6297).

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A Sharp I.F. Amplifier For 'Phone Or C.W.

The amplifier shown in Figs. 5-60 and 5-61 is designed to follow any receiver i.f. amplifier in the range around 455 kc., to give additional selectivity to the receiving system. For c.w. reception, ten circuits tuned to 50 kc. give a characteristic with excellent skirt selectivity, as indicated by a bandwidth of only 1900 cycles at −60 db. (Compare this with Figs. 5-1 and 5-22.) However, the amplifier is about 450 cycles wide at −6 db., so signals do not "ring" or become difficult to tune. For 'phone reception, some of the circuits are detuned (by throwing a switch) to give a "stagger-tuned" amplifier that has a bandwidth sufficient for reception of one sideband. However, since a majority of the circuits are still tuned to 50 kc., the resultant characteristic has greater gain at 50 kc. than at any other, and by tuning so that the heterodyned carrier falls at 50 kc., "exalted-carrier reception" is obtained. The useful bandwidth for 'phone reception is about 2300 cycles, so some high-audio-frequency response is lost, but the gain in intelligibility in crowded bands more than makes up for it. The bandwidth at −60 db. is 4000 cycles in the 'phone condition.

The complete circuit of the amplifier is shown in Fig. 5-59. Receiver output at 455 kc. is fed into the 6BE6 i.f. converter, where the crystal-controlled oscillator portion can be set either 50 kc. higher or lower, to use the familiar selectable-sideband principle.[1] If the receiver i.f. were something other than 455 kc., the choice of crystals would be different, of course. Two 6BJ6 i.f. amplifier tubes are coupled by eight Millen 50-kc. high-*Q* tuned circuits, *TC*$_1$–*TC*$_8$, with provision through *S*$_2$ for switching the tuning of four of these circuits by cutting in series condensers *C*$_5$, *C*$_8$, *C*$_{12}$ and *C*$_{16}$. The second 6BJ6 stage is coupled to a 6BE6 a.f. converter used for c.w. and s.s.b. suppressed-carrier reception, and also to another 6BJ6 i.f. amplifier. This i.f. amplifier feeds a diode rectifier for a.m. reception and also has a.v.c. voltage applied to its grid, to obtain some a.v.c. action and to give a logarithmic S-meter action. The switch *S*$_3$ selects output from one of the two detectors and also turns the b.f.o. on or off. The rest of the circuit includes a filter following *S*$_{3B}$ to keep 50-kc. energy out of the audio amplifier, and a meter amplifier for the S-meter. The i.f. gain is controlled through *R*$_7$, and the audio volume through *R*$_{25}$. The i.f. gain control is important, because the gain of the amplifier is much higher in the narrow-band (c.w.) condition than in the stagger-tuned ('phone) arrangement, and the gain setting must be changed when switch *S*$_2$ is thrown from one position to the other.

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[1] McLaughlin, "Exit Heterodyne QRM," *QST*, October, 1947.

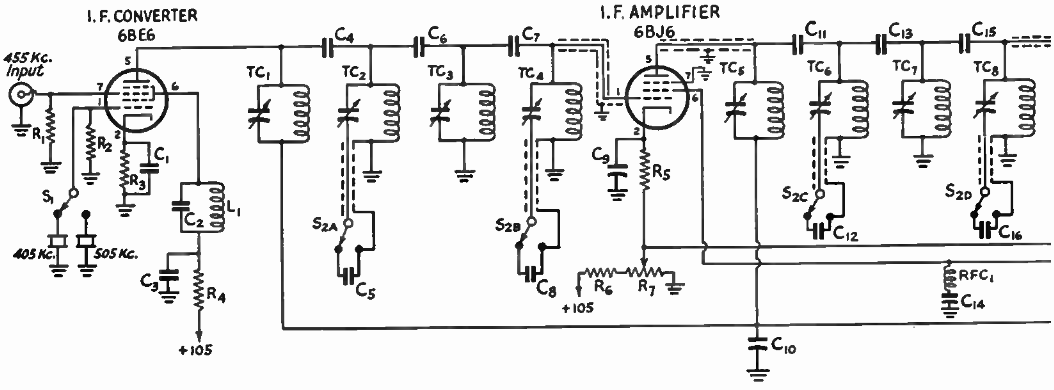

Fig. 5-59 — Wiring diagram of the selective 50-kc. i.f. amplifier.

C$_1$, C$_3$, C$_{29}$, C$_{30}$ — 0.01-μfd. ceramic disc.
C$_2$ — 47-μμfd. mica.
C$_4$, C$_6$, C$_7$, C$_{11}$, C$_{13}$, C$_{15}$, C$_{23}$ — 4.7-μμfd. mica.
C$_5$, C$_{12}$ — 0.0078-μfd. mica (0.0068 and 0.001 in parallel).
C$_8$, C$_{16}$ — 0.0068-μfd. mica.
C$_9$, C$_{17}$, C$_{22}$, C$_{27}$ — 0.1-μfd. 200-volt paper.
C$_{10}$, C$_{14}$ — 1.0-μfd. 400-volt paper.
C$_{18}$ — 100-μμfd. midget variable.
C$_{19}$, C$_{20}$, C$_{26}$, C$_{33}$, C$_{34}$, C$_{36}$ — 470-μμfd. mica.
C$_{21}$, C$_{24}$, C$_{31}$ — 0.001-μfd. mica.
C$_{25}$ — 100-μμfd. mica.
C$_{28}$ — 20-μfd. 25-volt electrolytic.
C$_{32}$ — 220-μμfd. mica.
C$_{35}$, C$_{38}$, C$_{41}$ — 0.01-μfd. 400-volt paper.
C$_{37}$ — 0.002-μfd. 400-volt paper.
C$_{39}$, C$_{42}$ — 10-μfd. 25-volt electrolytic.
C$_{40}$ — 8-μfd. 450-volt electrolytic.
R$_1$, R$_{21}$, R$_{33}$ — 0.47 megohm.
R$_2$, R$_{31}$ — 0.1 megohm.
R$_3$ — 470 ohms.
R$_4$, R$_{13}$ — 22,000 ohms.
R$_5$, R$_8$, R$_{10}$ — 100 ohms.
R$_6$ — 33,000 ohms, 2 watts.
R$_7$ — 5000-ohm wire-wound potentiometer.
R$_9$ — 2200 ohms, 1 watt.
R$_{10}$ — 2200 ohms.
R$_{11}$ — 33,000 ohms.
R$_{12}$ — 56,000 ohms.
R$_{14}$ — 330 ohms.
R$_{15}$, R$_{22}$ — 1.0 megohm.
R$_{17}$, R$_{18}$, R$_{32}$ — 4700 ohms.
R$_{19}$ — 68,000 ohms.
R$_{20}$, R$_{23}$ — 47,000 ohms.
R$_{24}$ — 1000 ohms.
R$_{25}$ — 0.25-megohm volume control.

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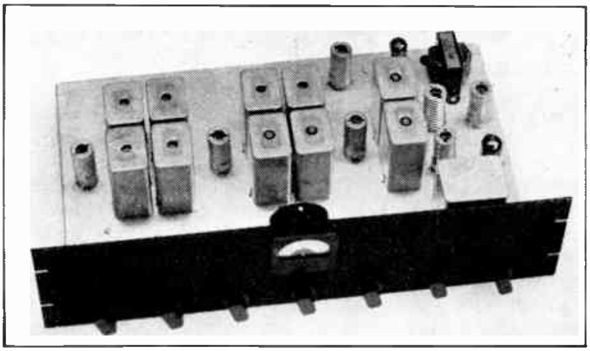

Fig. 5-60 — The selective 'phone and c.w. 50-kc. i.f. amplifier connects to the i.f. of a regular receiver, in the manner of the familiar Q5-er. It is built on a 10 x 17 x 2-inch chassis, with a 5¼-inch high panel. The plate circuit filter, C36, R23, RFC3, C37 and C38 are mounted above the chassis in the shield can near the right front.

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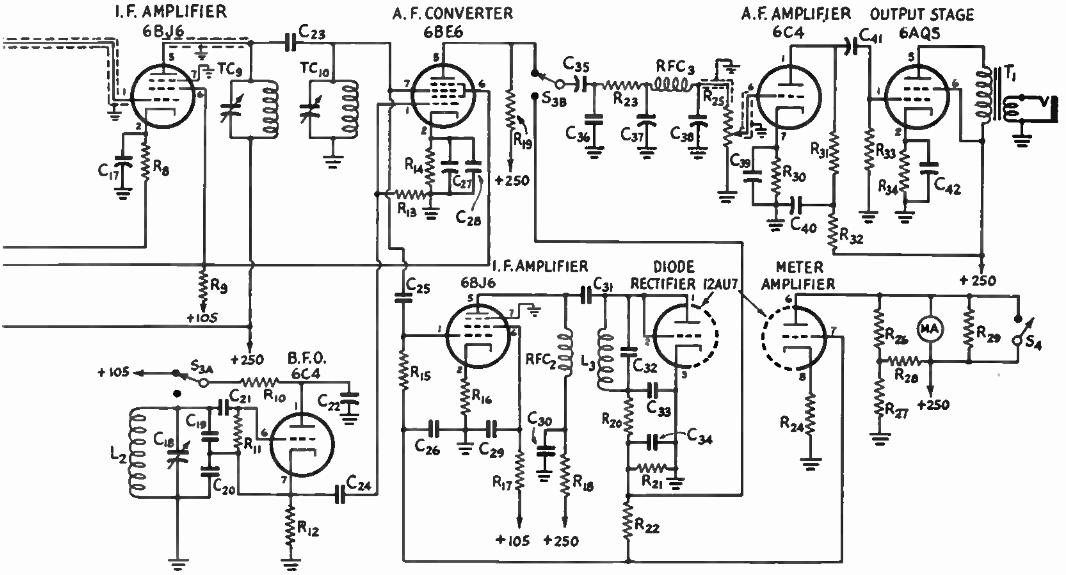
In an amplifier like this, over-all feed-back must be eliminated, and this calls for proper by-pass condensers in the common screen and plate leads. The choke, RFC_1, is included to series resonate with C_{14} at 50 kc. and bring the common screen circuit down to ground potential without using a by-pass condenser larger than 1 μfd. Everything else in the circuit is conventional and familiar to anyone acquainted with i.f. amplifier practice. The switch, S_4, for shorting out the S-meter, is an elaboration except for someone who doesn't like to see the meter swing too far on occasion.

Constructional details will, of course, vary with the builder and his particular station layout. The only point to watch in any modification is to be sure not to double the amplifier string back on itself, which might encourage feedback.

Adjustment

The first step in aligning the amplifier, after the wiring has been checked, is to tune the circuits to 50 kc. approximately. For a signal source, couple from the receiver at 455 kc. and feed it into the input of the amplifier. Tune in a steady signal (freqmeter or b.c. carrier) and align the tuned circuits TC_{10} back through TC_1, with S_2 set to short out the condensers. When the circuits have been aligned, as indicated by the S-meter, switch S_1 to the other crystal and see if the signal is still peaked. It probably won't be, and you now have to "split the difference" and realign the circuits until a signal peaked at one setting of S_1 will be peaked at the other. If the exact



R_{26} — 220 ohms.
R_{27} — 47,000 ohms, 2 watts.
R_{28} — 390 ohms.
R_{29} — 47 ohms.
R_{30} — 3900 ohms.
R_{34} — 270 ohms.
L_1 — 750 uh. (National R-33).
L_2 — 40 mh. (Millen 34240).
L_3 — 80 mh. (Millen 34280).
MA — 0.1 milliammeter.
RFC_1 — 67 t. No. 30 enam. close-wound on ¼-inch diam. form (1-megohm resistor).
RFC_2 — 80 mh. (Millen 34280), mounted at right angles to L_3.
RFC_3 — 125 mh. (Millen 34000-2 removed from can).
S_1, S_3, S_4 — D.p.d.t. rotary switch (Mallory 3222J); one pole only used on S_1 and S_4.
S_2 — 4-pole rotary switch (Mallory 3242J).
T_1 — Audio output transformer (Merit A-2902).
TC_1-TC_{10} — 50-kc. tuned circuit (Millen 63650).

frequencies of the crystals are known, and you have an accurate signal generator in the 50-kc. range, align the circuits the first time on half the frequency difference between the two crystals. It may sound complicated, but it isn't. The experimental method shouldn't require more than four or five tries.

Next check the b.f.o. tuning range by switching it on and watching the S-meter as C_{13} is tuned through its range. The b.f.o. couples some energy into the S-meter circuit, and as C_{13} is tuned through you will find a peak. Add or subtract fixed capacity across the circuit until the peak occurs with C_{13} at about half capacity. Tuning the b.f.o. through its range should show an even rise and fall in the S-meter circuit — any sudden jump indicates some regeneration in the amplifier. Switch off the b.f.o., tune in a signal as indicated on the S-meter, and jump a 1-μfd. condenser across the B+ and screen leads at various points throughout the amplifier. Any change in S-meter reading indicates regeneration in the amplifier, and it means that more by-passing, or a modification of RFC_1, is required. In adjusting RFC_1, wind it large and remove turns two at a time, with a constant signal through the amplifier. When the signal comes down to a minimum and then starts back up, rewind the choke to the turns that gave the minimum. The amplifier may be slightly regenerative at a high gain setting in the "sharp" condition. However, this gain is well beyond anything you would ever use, and at normal settings the amplifier should be perfectly stable. Also check the S-meter reading with the b.f.o. on (no signal coming through) to see if there are any points along the +105-volt line where extra by-passing will cut down the b.f.o. leakage.

At about this point in the checking it is wise to try a little more capacity across C_{32}, since this fixed condenser tunes the coupling circuit between the 6BJ6 amplifier and the diode. With a steady signal coming through, hold a 10- or 20-μμfd. condenser across C_{32}. If the S-meter reading goes up, C_{32} is too small, and if the reading goes down the condenser is too large. The tuning isn't very critical however, and there is plenty of signal to spare at this point.

Now switch S_2 to the other position and tune across an unmodulated signal, with the b.f.o. off and R_7 set for maximum gain. The meter reading will show two peaks, one of the same frequency as in the "sharp" position and one new one. Experiment with the tuning of TC_9 and TC_{10} until the new peak is not so sharp and has somewhat less amplitude than the original. You will find that you can change it considerably, so make your changes in small steps. Set it up by cut and try, using a b.c. station as your signal and adjusting for good intelligibility. It probably won't be necessary to touch C_5, C_8, C_{12} and C_{16}, unless these condensers are incorrectly marked and thus far off the nominal value. Their value, of course, affects the position of the lower hump in the selectivity characteristic.

When tuning in a 'phone signal, watch the S-meter and set the carrier at the maximum meter reading, learning not to be fooled by the slight hump 2000 cycles off this point. If there is a heterodyne on the signal, flip S_1 to the position giving the least interference.

In operation, set the i.f. gain controls in both receiver and 50-kc. i.f. where there is no chance for overload, with the audio adjusted for comfortable listening. On c.w. the S-meter may kick a little on a loud signal, and on 'phone it will get up to half scale on a good signal. This calls for some juggling of the gain controls as you tune across a band. When throwing S_3 from one position to the other, it is advisable to reduce the audio volume to zero first, or else the audio tube takes quite a licking because of the difference in potential in the two points on switch S_{3B}.

You may have to realign your receiver slightly to use this amplifier, or else get a pair of crystals that match your receiver. If, for example, the crystal in your present receiver is on 450 kc., and hence the i.f. is aligned there, the 405/505-kc. crystal combination won't be right. What you need is a 400- and a 500-kc. crystal. But since the crystal filter won't be used much any more, it is easier to switch it out of the circuit and realign the receiver i.f. amplifier. Tune in a signal on the 50-kc. amplifier and retune the receiver i.f. transformers for maximum signal, and that's all there is to it.

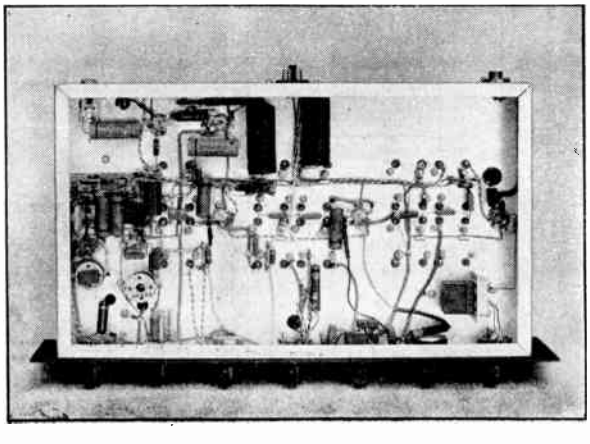


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Fig. 5-61 — Underneath the chassis of the 50-kc. amplifier. Leads to the audio volume control, to the grids and plates of the amplifier tubes, and to the selectivity switch, are all run in shielded wire.

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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.

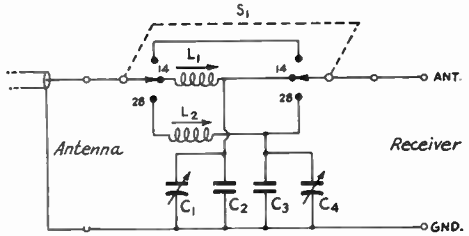

Fig. 5-62 — **Wiring diagram of the "L"-section matching network.**

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 ⅝ 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.

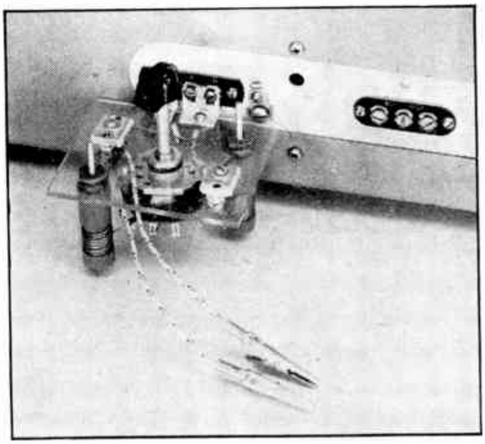

Fig. 5-63 — **The "L"-section coupler mounted on the antenna and ground binding posts of a communications receiver.**

An "L"-section matching coupler for 14 and 28 Mc. is shown in Fig. 5-64. 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. 5-62, 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.

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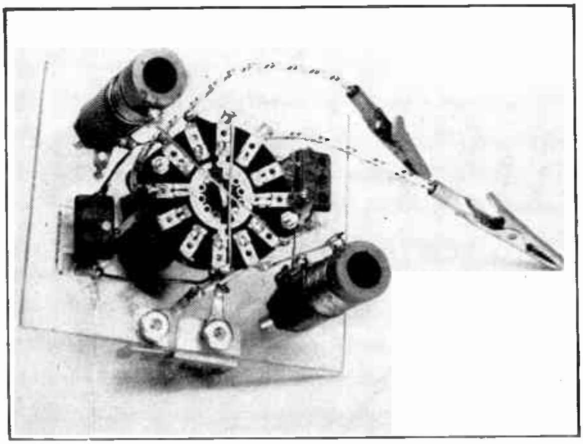

Fig. 5-64 — **An "L"-section matching network for coupling the receiver to coaxial line. It is designed for use between 50- or 75-ohm line and a receiver of 300 to 400 ohms input impedance.**

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CHAPTER 6

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.

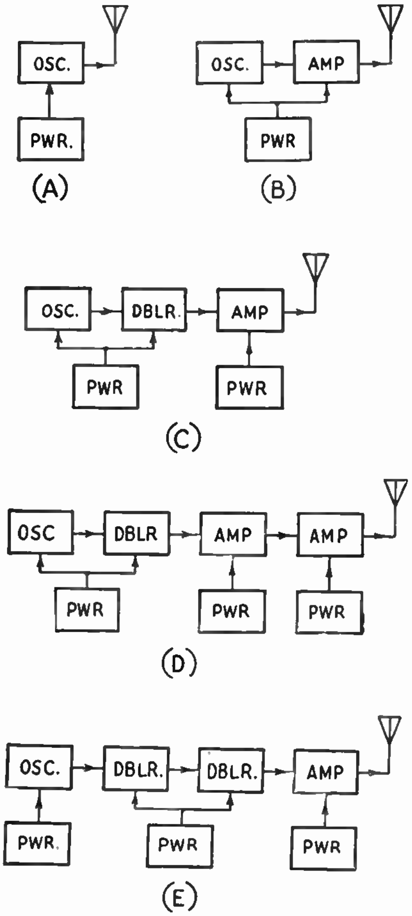

Fig. 6-1 — Block diagrams showing typical combinations of oscillator and amplifiers and power-supply arrangements for transmitters. A wide selection is possible, depending upon the number of bands in which operation is desired and the power output.

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 megacyle) by a quartz crystal. The frequency is determined almost entirely by the thickness of the crystal. Other constants in the circuit have relatively little effect. The frequency of a self-controlled or variable-frequency oscillator (VFO) is determined principally by the values of inductance and capacitance which make up the oscillator tank circuit.

The disadvantage of the crystal type of oscillator is that a different crystal must be used for each frequency desired (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

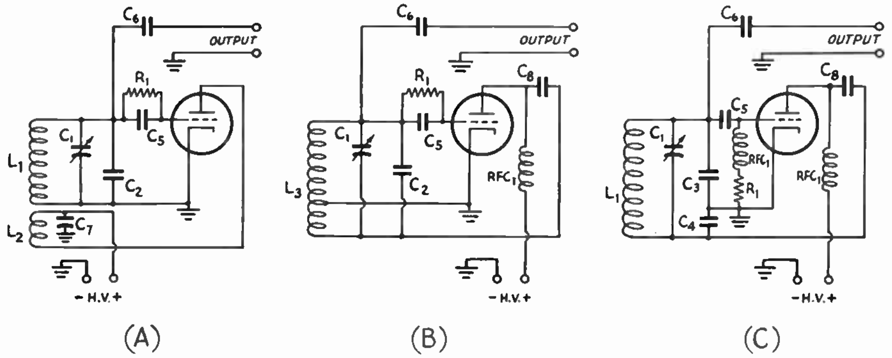

Fig. 6-2 — Typical simple VFO circuits. A — Tickler feed-back. B — Hartley. C — Colpitts. Low-μ triodes, such as the 6C5 or 6J5, are preferable. Approximately appropriate values for the 3.5-Mc. band are as follows:

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 — 0.01-μfd. disk ceramic.
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 to isolate the two.

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.

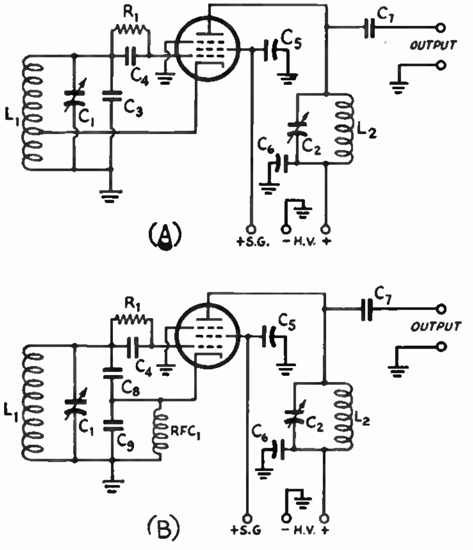

Fig. 6-3 — ECO circuits. A — Hartley. B — Colpitts. Approximate values are as follows:

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 — 0.01-μfd. disk ceramic.
C_6 — Plate by-pass — 0.01-μfd. disk ceramic.
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.

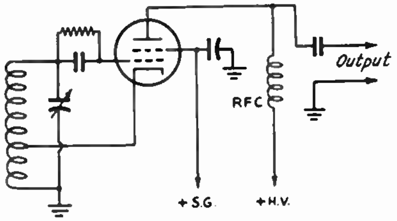

Fig. 6-4 — **ECO with an r.f. choke replacing the output tank circuit for the purpose of reducing reaction on the oscillator portion of the circuit.**

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.

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 the chapter treating keying, 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.

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.

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 capacitance changes a small percentage of the total circuit capacitance. At 3.5 Mc., a tank capacitance of 500 to 1000 μμfd. 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. 6-5A 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

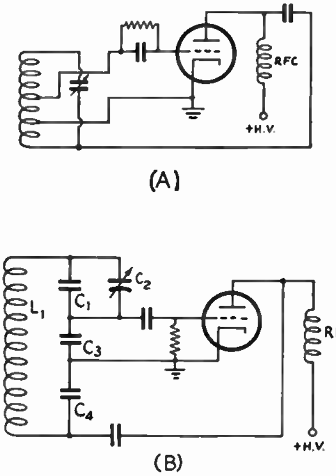

Fig. 6-5 — As an alternative to the use of a high-C tank circuit, oscillator tubes sometimes are connected across only a portion of the tank circuit to increase the Q. In the Hartley circuit of A, the grid and plate connections are made to taps instead of to the ends of the coil. In the Colpitts circuit 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.

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.

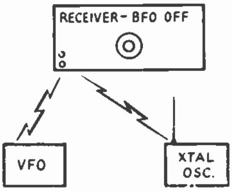

Fig. 6-6 — Set-up for checking VFO stability. The receiver should be tuned preferably to a harmonic of the VFO frequency. The crystal oscillator may operate somewhere in the band in which the VFO is operating. The receiver b.f.o. should be turned off.

In the circuit of Fig. 6-5A, 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. 6-5B, 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. 6-6. (See "Crystal Oscillators," this chapter.) 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.

● SIMPLE CIRCUITS

The basic crystal-controlled oscillator circuits are shown in Fig. 6-7. 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. 6-7A 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. 6-8. 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 greatest. As the tuning capacitance is decreased further, the plate current will rise gradually to point *B* where it will jump to a higher value, indicating that oscillation has ceased. For maximum frequency stability the circuit should be tuned in the region *D-E*.

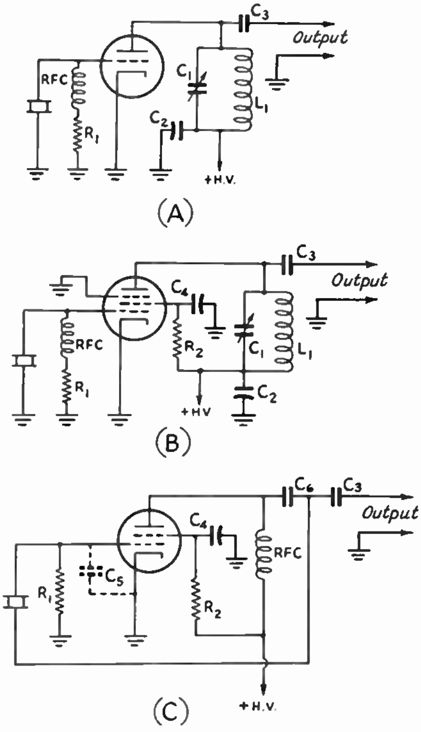

Fig. 6-7 — Simple crystal-oscillator circuits. A — Triode. B — Tetrode or pentode. C — Tetrode Pierce. Approximate values are as follows:

C1 — Tank condenser — 100-μμfd. variable.
C2 — Plate by-pass — 0.01-μfd. disk ceramic.
C3 — Output coupling condenser — 100 μμfd. or less, mica.
C4 — Screen by-pass — 0.01-μfd. disk ceramic.
C5 — Feed-back condenser — 50 to 100 μμfd.
C6 — Plate blocking condenser — 0.001-μfd. mica.
R1 — Grid leak — 50,000 ohms.
R2 — Screen voltage-dropping resistor — 25,000 to 50,000 ohms.
L1 — Tank coil — 22 μhy. for 3.5 Mc.; 7.5 μhy. for 7 Mc.
RFC — Parallel-feed r.f. choke — 2.5 mh.

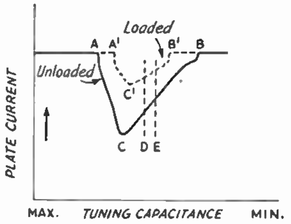

Fig. 6-8 — General tuning characteristic of triode, tetrode or pentode crystal oscillators. As the capacitance of the plate tank condenser is decreased from maximum, oscillation will start suddenly, the plate current dropping abruptly, as shown at point *A*. As the capacitance is decreased further, the plate current will dip to a minimum, *C*, and then gradually rise to point *B* where an abrupt rise in plate current will indicate that oscillation has ceased. Maximum output will be obtained at point *C*, but the oscillator should be adjusted for operation in the *D-E* region for best frequency stability.

When the oscillator is loaded, the characteristic is similar (see dashed curve in Fig. 6-8), but the minimum plate-current dip is much less pronounced and the range of plate tuning over which the circuit will oscillate becomes less as the loading is increased.

With triode, tetrode or pentode circuits, the feed-back may be adjusted by the tuning of the plate tank circuit, as in the t.g.t.p. circuit. However, it is safer to limit the amount of power fed back from the plate circuit by restricting the plate voltage so that the crystal limitation cannot be exceeded inadvertently during normal tuning procedure. With the large prewar-type crystals, triode crystal oscillators may be operated with proper adjustment at plate voltages as high as 200 or 250, but the voltage should be reduced to 100 to 150 for the new-type smaller crystals. Low- or medium-μ tubes are preferable when triodes are used. Beam tetrodes or pentodes, with their high power-sensitivity and reduced grid-plate capacitance, require less voltage across the crystal than a triode for the same amount of output. With screen-grid tubes the larger-type crystals can be operated with plate voltages of 300 or 400 with power output up to 10 or 15 watts if required. The smaller crystals may take plate voltages up to 250 or 300 before showing marked drift in frequency. However, as stated previously, it is always advisable to limit the input to the oscillator and depend upon amplifiers for the desired power.

With the triode, tetrode and pentode crystal-oscillator circuits shown in Fig. 6-7, excitation, and therefore r.f. voltage across the crystal, is greatest when the oscillator is unloaded and it is under this condition that danger to the crystal is greatest.

Pierce Oscillator

The circuit shown in Fig. 6-7C is known as the Pierce circuit. It is the equivalent of the ultraudion variation of the Colpitts in which the grid-cathode and plate-cathode capacitances of the tube form the capacitive divider. The crystal replaces the single tuned circuit and thus this oscillator requires no tuning adjustment and will work without change in values over a wide range of crystal frequencies. Since excitation otherwise is not adjustable, except by change in plate voltage, the condenser C_5 sometimes is required to obtain satisfactory operation. Less power is obtainable from the Pierce circuit than from the preceding oscillators because the crystal is directly in the power-delivering circuit which limits the r.f. voltage that may be developed without danger to the crystal. Triodes also may be used in this circuit.

COMBINATION CRYSTAL OSCILLATORS

Tri-Tet Circuit

Fig. 6-9 shows three crystal-oscillator circuits which operate on principles similar to those of the ECO. Circuits such as these have the additional advantage that they are invariably found to key more reliably than the simple triode or tetrode circuits, and do not incur the considerable loss in efficiency sometimes involved in detuning the plate circuit far to the high-frequency side of resonance for reliable crystal starting under load.

The extent to which the output-plate circuit reacts on the oscillator portion of the circuit, and the output-circuit tuning characteristics, are influenced to a considerable degree by the effectiveness of the screening of the tube selected. Well-screened tubes always are preferable from the standpoints of both isolation and safety to the crystal.

Fig. 6-9A shows the Tri-tet circuit. The oscillator portion is equivalent to that of a triode crystal oscillator, with the screen serving as the "plate" and the ground point being shifted from the cathode to the "plate." Power is taken from a separate output-plate tank circuit. Since the output-plate circuit returns to cathode through the L_1C_1 tank circuit, the plate contributes to the feed-back to a certain extent. In addition, the screen-to-control-grid capacitance is greater than the corresponding plate-grid capacitance of most triodes. Therefore, L_1C_1 should always be tuned well to the high-frequency side of the crystal frequency to prevent excessive feed-back and consequent unnecessary crystal heating. In respect to heating of the crystal, limitation of screen voltage is of greater importance than plate voltage.

As with the ECO, harmonic output may be obtained by tuning the output tank circuit, L_2C_2, to the desired multiple of the crystal frequency.

When operating the Tri-tet circuit at the crystal frequency, L_1C_1 should be tuned no closer to the crystal frequency than is necessary to make the circuit oscillate without output-plate voltage applied. With poorly-screened tubes, it may be advisable to short-circuit L_1C_1 when operating at the crystal

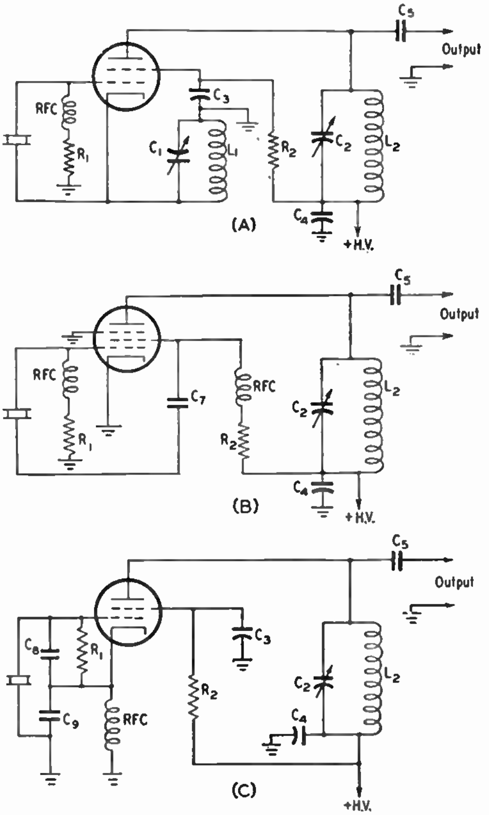

Fig. 6-9 — Crystal-controlled amplifying-oscillator circuits. A — Tri-tet. B and C — Modified Pierce. Approximate values are as follows:

C_1 — Cathode-tank tuning condenser — 100-μμfd.
C_2 — Output-tank tuning condenser — 100-μμfd.
C_3 — Screen by-pass condenser ("plate" grounding) — 0.01-μfd. disk ceramic.
C_4 — Plate by-pass — 0.01-μfd. disk ceramic.
C_5 — Output coupling condenser — 100 μμfd. or less.
C_6 — Feed-back control condenser — 100-μμfd. variable.
C_7 — Parallel-feed blocking condenser — 0.001-μfd.
C_8 — Feed-back-adjustment condenser — 10 to 15 μμfd.
C_9 — Feed-back-adjustment condenser — 150 to 200 μμfd.
R_1 — Grid leak — 50,000 to 150,000 ohms.
R_2 — Screen voltage-dropping resistor — 25,000 to 100,000 ohms.
L_1C_1 — Tuned well above crystal frequency (see text).
L_2C_2 — Tuned to crystal frequency or desired harmonic.
RFC — Parallel-feed r.f. choke — 2.5 mh.

fundamental frequency, reverting to the tetrode circuit as a measure of safety to the crystal.

With well-screened tubes, such as the 6SK7, 6AG7 or 802, the output-plate tuning characteristic is like that shown in Fig. 6-10 at the fundamental as well as at the harmonics, and the circuit will continue to oscillate regardless of the tuning of the output circuit. However, with poorly-screened tubes, such as the 6V6, 6F6 or 6L6, the circuit may stop oscillating abruptly when the output circuit is tuned to a frequency lower than the crystal frequency, more in the manner of a simple triode or tetrode oscillator. With well-screened tubes, feed-back is at a minimum when the output circuit is unloaded, the excitation increasing as load is increased. This characteristic is opposite to that of the triode or tetrode crystal oscillator.

Modified Pierce Circuits

Figs. 6-9B and 6-9C show similar extensions of the Pierce or Colpitts-type circuit. In B, the oscillator portion of the circuit is the triode Pierce with the cathode grounded and the screen serving as the "plate." In this arrangement the suppressor provides screening against capacitive coupling. Pentodes, such as the 6SK7, 6AG7 and 802, are suitable for this circuit.

Variation in screen voltage provides the primary means of adjusting feed-back, although it may also be changed by adding capacitance between either grid and cathode or screen and cathode, as found necessary. Harmonic output may be obtained by tuning the output circuit to a multiple of the crystal frequency.

The circuit of Fig. 6-9C is identical, except that the operation is less dependent upon the relative electrode capacitances of the tube. C_8 and C_9 are connected in parallel respectively with the grid-cathode and "plate"-cathode capacitances of the tube. This is a more flexible arrangement than that of B, since the amount of feed-back is now a function of the relative capacitances of the two condensers, rather than of the tube capacitances. One or both of the condensers may be made variable to provide a means of adjusting the feed-back over a wide range.

With some adjustments of feed-back, it may be found that the crystal will stop oscillating when the output circuit is tuned exactly to 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. 6-9 are suggested, but they may have to be modified somewhat, depending upon the type of tube used.

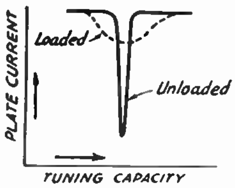

Fig. 6-10 — Plate-tuning characteristic for Tri-tet, grid-plate and modified Pierce crystal-oscillator circuits, using well-screened tubes, for loaded (dashed line) and unloaded (solid line) conditions. In this case the output-plate circuit may be tuned accurately to the minimum plate-current dip for maximum output.

QUARTZ 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 the frequencies of 7002 kc., 14,004 kc., 28,008 kc. etc.

The characteristics of a crystal — particularly in the thickness-frequency and temperature-frequency relationships — depend upon the plane in which the crystal plate is cut from the natural quartz block. While other cuts are useful in certain other applications, those for amateur transmitters invariably are of either the "AT" or "BT" types. Their respective temperature characteristics are as follows:

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:

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

where $f_{\mathrm{Mc.}}$ 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:

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.

Regrinding Crystals

Because crystals near any desired frequency can be purchased reasonably these days, it is not profitable for the amateur to cut and grind his own blanks. However, frequently it may be desirable to make a limited increase in the frequency of a crystal at hand. Indispensable requirements are a piece of plate glass, a good micrometer, supplies of Size 800 aluminum oxide for light grinding, and Size 400 silicon carbide for coarse grinding, and a test oscillator. The coarse grinding compound may be desirable only if the required change exceeds 40 or 50 kc. at 7 Mc. or half of this at 3.5 Mc. Even much larger changes can be made with the finer abrasive — it simply takes somewhat longer. A test oscillator of the regenerative type, such as the one shown in Fig. 6-9C, is preferred because it oscillates more readily and therefore permits frequency checks at times when the activity of the crystal may be subnormal. The oscillator should be equipped with a grid-current milliammeter, preferably one with a 0.5-ma. scale. The grid current should be checked first with the crystal to be reground, and preferably with several others, to obtain an average of the grid current to be expected for normal crystal activity. Then this figure can be used as a standard of comparison, less grid current indicating lower activity.

In grinding a crystal to a new frequency, the most important factor in respect to activity is that of maintaining the proper surface contour. When properly ground, the crystal is thicker in the center than at the edges. The difference in thickness should vary from about 0.001 inch for a 3.5-Mc. crystal ½ inch square to about 0.00015 inch for a 7-Mc. crystal. The micrometer is used chiefly in checking the contour.

The grinding compound should be sprinkled on the glass plate and moistened with water to make a very thin paste. One side of the crystal should be marked with a pencil and *all* of the grinding should be done on the *opposite* side. Thus the marked side is always preserved for reference in checking with the micrometer. The crystal should be swirled around in the grinding mixture in a motion describing figure-eight paths. The location of the path should be changed frequently to another part of the glass plate so that the plate will be worn evenly. Light pressure with the finger on a corner of the crystal should be used. Make three or four "8's" to each of the corners in succession and then repeat. Use lighter pressure and make fewer "8's" as the desired frequency is approached. It is a good idea to mark the reference side with a character — such as an **F** — which can be read properly in but one position. This will help to identify each of the four corners. After each corner has had a turn, the identifying mark should be back to the original starting position, of course.

If a calibrated receiver is available, it can be used to keep a continuous check on the frequency as the crystal is being ground. Place a sheet of tinfoil or metal under the plate glass and connect it to the antenna terminal of the receiver. Then as the crystal is being ground, it will produce a hiss in the receiver that peaks close to the crystal frequency. To be safe, however, it is advisable to limit the use of this method of checking to within 20 kc. of the desired frequency at 7 Mc. Then if it is found that the activity is not up to normal, the contour can be corrected without overshooting the desired frequency. It might be mentioned also that this method may not be usable at night when outside signals may make it difficult to spot the crystal frequency with any great accuracy.

The crystal should be thoroughly cleaned of grinding compound and other matter before using the micrometer or checking in the test oscillator, of course. Use soap, warm water and a tooth brush, and dry with a lintless cloth or tissue. Handle the crystal by the edges only after cleaning.

In measuring the corners, the apex of the corner should be placed at the center of the micrometer jaw. Since accurate micrometer readings are difficult when they differ so slightly, it is sometimes helpful to judge by "feel." Set the micrometer lightly to the center of the crystal and then move the crystal around. In no case should there be any binding or sticking at the edges with the micrometer set at the same position.

The crystal should be checked in the test oscillator more frequently as the desired frequency is approached, both for frequency and activity. Up to this point it should be sufficient merely to place the crystal in the holder and squeeze the assembly together with the fingers while making the check. Make sure, however, that the outside plate of the holder is connected to the "ground" side of the circuit. When close to the desired frequency, the checks should be made with the holder completely assembled. If the activity is found to have dropped off, pay particular attention to the corners and bring the activity back to normal before the final fine adjustment to frequency. In some cases, beveling or grinding one or more of the edges may be necessary to restore activity, but this should be done with caution and only in case the contour has first been corrected. If the crystal width and length dimensions are responsible for low activity, a few light swipes on the edges should be sufficient to show an improvement in activity.

Lowering Frequency

If a crystal has accidentally been ground down too far, or if it is desired to lower slightly the frequency of any other crystal, this can often be done by loading the crystal. Loading, however, may reduce the crystal activity if it is carried too far. With a good active crystal, it should be possible to decrease the frequency as much as one per cent — 35 kc. for a 3500-kc. crystal. Cold soft solder rubbed into the crystal surface is suitable. The solder should be applied gradually while the frequency and activity are checked. Start off by marking a circle about ¼ inch in diameter at the center of the crystal and use this as a boundary for additional applications of the solder. The loading should be applied to both surfaces as equally as possible.

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.

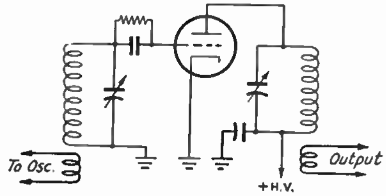

Fig. 6-11 — Fundamental r.f. power-amplifier circuit. Means must be provided to prevent oscillation since the circuit is the same as that for a t.g.t.p. oscillator. See text for discussion.

Fig. 6-11 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.

The most commonly-used circuits for this purpose are shown in Fig. 6-12. Amplifiers using these systems of neutralization are

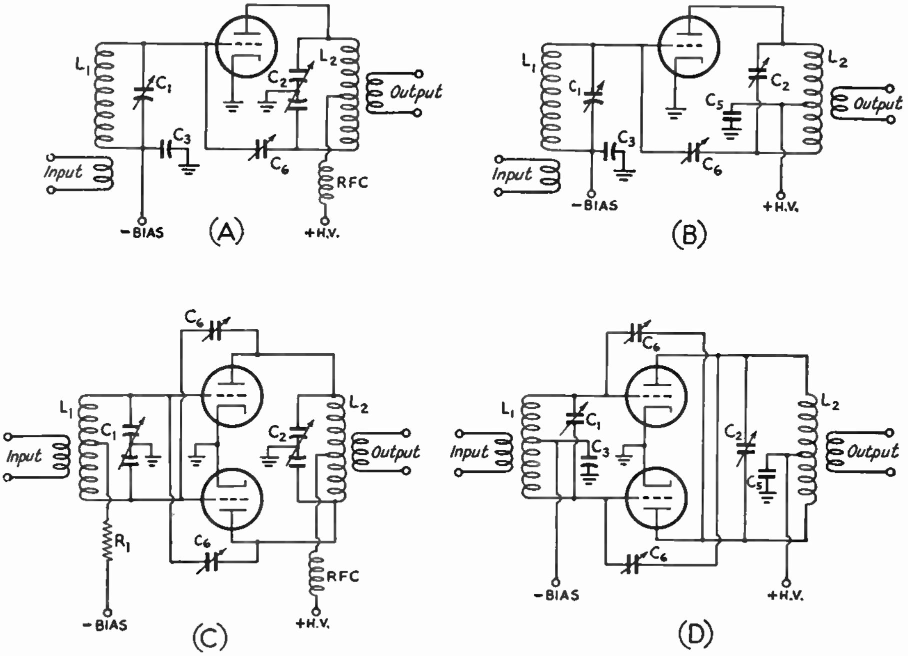

Fig. 6-12 — Neutralized-triode amplifier circuits. A — Single tube with capacitive balance. B — Single tube with inductive balance. C and D show corresponding push-pull arrangements.

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.

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

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.

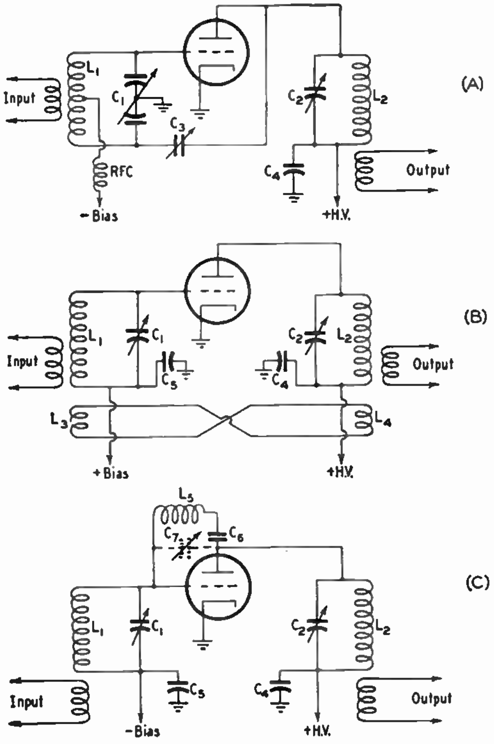

Fig. 6-13 — Additional, but less commonly-used neutralizing circuits. A — Grid neutralizing. B — Link neutralizing. C — Inductive neutralization.

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.
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 — usually a single turn is sufficient.
L_5 — Neutralizing trap coil — to tune to operating frequency with C_7 and grid-plate capacitance of tube. (See text.)

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. 6-12A, the division of voltage across the tank circuit is dependent upon the ratio of the capacitances of the two sections of the tank condenser. Since these capacitances are equal in a split-stator condenser, the voltages at the ends of the tank circuit in respect to the cathode, which is connected to the center of the tank circuit through ground, are equal. Therefore the neutralizing voltage is the same as the feed-back voltage when the capacitance of the neutralizing condenser is equal to the grid-plate capacitance of the tube, including socket and other external stray capacitances across the elements.

In Fig. 6-12B, the voltage division for neutralization is dependent upon the ratio of inductances in the two sections of the coil. The coil usually is tapped at the center to give equal voltages at the ends of the tank circuit.

Push-Pull Triode Circuits

Fig. 6-12C 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.

In Fig. 6-12C, 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. 6-13. The circuit of Fig. 6-13A is similar to that of Fig. 6-12A, except that the voltage division takes place in the grid circuit instead of the plate circuit. Any voltage which may be fed back to the grid circuit through the grid-plate capacitance of the tube is divided in the grid tank circuit so that half appears at the grid, while the other half is fed, 180 degrees out of phase, back to the plate. In another similar version the grid tank coil, instead of the condenser, is used as the voltage divider, the circuit being comparable to Fig. 6-12B.

Link Neutralizing Circuit

The link neutralizing circuit of Fig. 6-13B 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. 6-13C consists merely of making the plate-grid capacitance of the tube part of a high-*Q* 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. Most complete neutralization requires that the *Q* of the tuned circuit be high. On the other hand, the higher the *Q* the narrower will be the band over which a single adjustment will hold.

All of the circuits of Fig. 6-13 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. 6-14. 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.

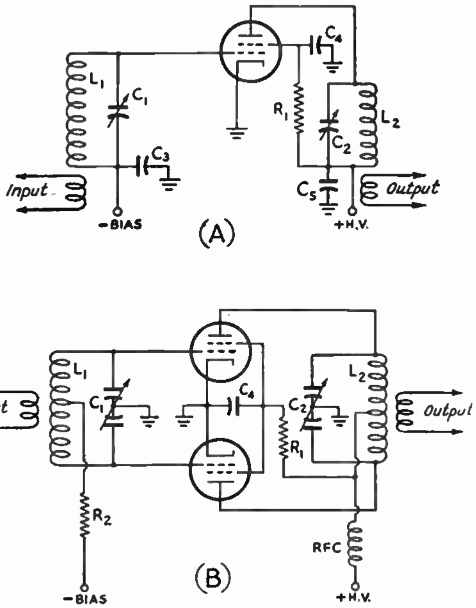

Fig. 6-14 — Screen-grid amplifier circuits. A — Single tube amplifier. B — Push-pull.

C_4 — Screen by-pass condenser — 0.001-μfd. mica.
R_1 — Screen voltage-dropping resistor.
R_2 — Grid-circuit isolating resistor — 100 ohms.
Other values same as Fig. 6-12.

Interstage Coupling Systems

Efficient transfer of power from the output tank circuit to the grid of a succeeding amplifier requires the use of a suitable correctly-adjusted coupling system. The essential requirement of such a system is that it should, so far as possible, transform the actual load impedance presented by the grid circuit of the following tube to the value of output load impedance recommended as optimum for the driving tube. (See section on tube operating factors, this chapter.)

● INDUCTIVE COUPLING METHODS

Link Coupling System

The inductive link coupling system, several examples of which are shown in Fig. 6-15, has a number of advantages. The two tuned circuits discriminate against harmonics and therefore this system is a primary means of reducing TVI, especially when used between the transmitter output stage and its driver. It facilitates coupling between balanced and unbalanced circuits and matching the optimum load impedance for the driver and the input impedance presented by the driven stage when they differ appreciably in value. Since the output capacitance of the driver stage and the input capacitance of the driven stage are across individual tank circuits, the system is preferred when circuit capacitance must be minimized. When properly adjusted, the link line can be of any length with negligible loss or radiation and therefore is useful when it becomes necessary or desirable to have considerable physical separation of driver and driven stages.

Fig. 6-15A 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.

The diagram of Fig. 6-15C shows the method applied in coupling the output of an unneutralized driver to a push-pull amplifier circuit, while D is the circuit to be used in coupling a neutralized or push-pull stage to another push-pull grid circuit.

This system may be considered either as simply a means of obtaining mutual inductance between the two tank coils or as a low-impedance transmission line. If the system is to be treated without regard to transmission-line effects, certain requirements must be met if sufficient coupling is to be obtained. The *Q* of each of the two tank circuits must usually be at least 10 and preferably 12 (see "Amplifier Design Considerations", this chapter). The link coils should have an equal number of turns and should be capable of being coupled tightly to the tank coils. The link line must not offer appreciable reactance at the operating frequency.

The first two requirements are not often difficult to meet, but the third requires special consideration unless the line is very short. Any appreciable reactance, inductive or capacitive, will in effect reduce the coupling. Coaxial cables especially have considerable capacitance for even short lengths and for this reason it may be more desirable to use a spaced line, such as twin lead, if the radiation can be tolerated.

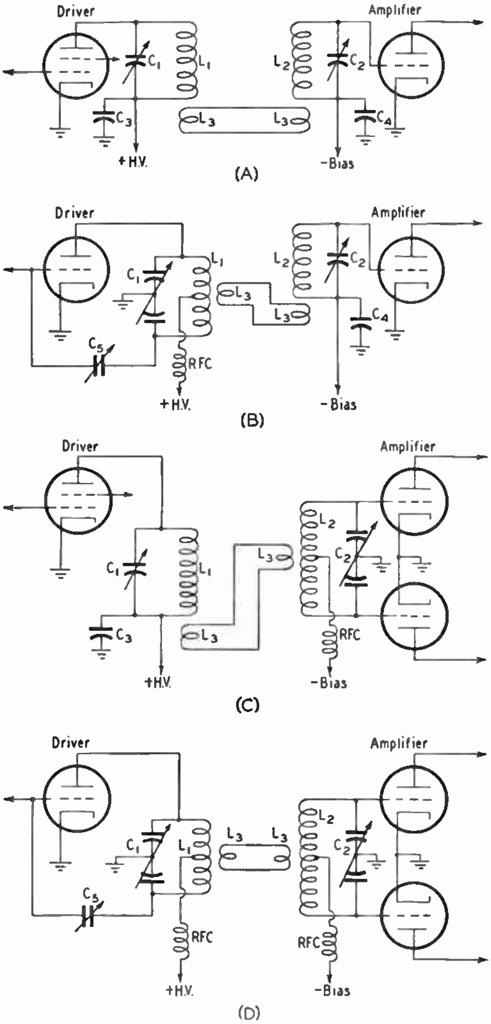

Fig. 6-15 — **Link coupling circuits. A — Unbalanced output to unbalanced input. B — Balanced output to unbalanced input. C — Unbalanced output to balanced input. D — Balanced output to balanced input.**

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.
RFC — R.f. choke.
C_1L_1 and C_2L_2 are always tuned to the same frequency.

The reactance of the line can be nullified only by making the link resonant. This may require changing the number of turns in the link coils, the length of the line, or the insertion of a tuning capacitance. The disadvantages of such a resonant link are obvious. The line losses increase because of the greater current, the voltage increase may be sufficient to cause a break-down of the cable and the added tuned circuit makes adjustment more critical with relatively small changes in frequency.

These troubles may not be encountered if the link line is kept very short for the highest frequency. A length of 5 feet or more may be tolerated at 3.5 Mc., but a length of a foot of coaxial line at 28 Mc. may be enough to cause serious effects on the functioning of the system.

Most of these difficulties can be avoided if the link is considered as a transmission line. However, a standing-wave indicator is required for proper adjustment. The procedure is described in detail in the chapter on transmission lines.

Coupling is adjusted by moving one or both of the link coils in respect to their respective tank coils.

Inductive and Bandpass Coupling

When negligible separation between stages is permissible, the link line may be omitted and the two tank coils coupled directly. Coupling is adjusted by changing the position of one coil in respect to the other. With the coils in close proximity, the amount of capacitive coupling introduced by the capacitance between coils may be greater than with link coupling where the coils usually have greater separation. This may be a consideration in reducing v.h.f. harmonics to a minimum.

If the *L/C* ratios in the two tank circuits are appropriate and the coupling is adjusted critically, it is possible to cover the width of any amateur band (up to and including 28 Mc.) with only a single initial adjustment of the two tuning condensers. In other words, the tuning is "broadbanded." To adjust the coupling, the two coils are first separated as much as possible, so that the tuning of one circuit will not disturb the tuning of the other too drastically. Each circuit is then tuned to resonance near the center of the band to be covered. Then the coupling is gradually increased until a characteristic similar to that shown in the solid line of Fig. 6-16 is obtained. Overcoupling is indicated by peaks near the ends of the band, with a dip at the center, while a peak at the center and drooping at the ends of the band are characteristic of undercoupling. Where the desired band cannot be covered satisfactorily, the *L/C* ratio should be increased. This may not be possible at 14 and 28 Mc. without taking special precautions to minimize circuit capacitances.

In the case of a single peak, care should be taken to make sure that the peak observed is in fact a single peak and not one of two peaks, the other lying outside the observable range. Such a situation may arise if the preceding circuits are

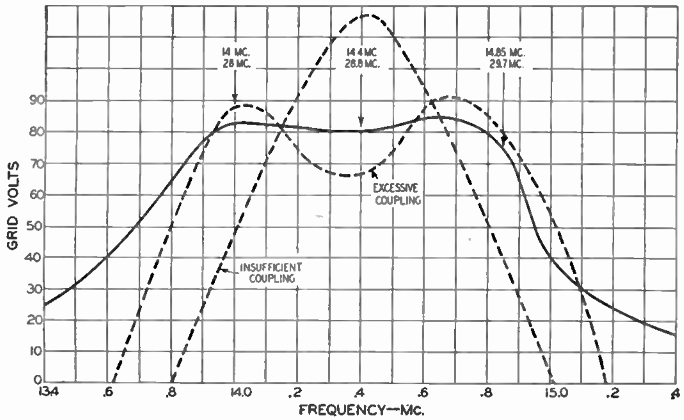

Fig. 6-16 — Typical bandpass-coupler characteristics. The solid curve shows the sort of performance possible with a correctly-adjusted coupler. The dashed-line curves are typical of incorrect coupling.

limited in frequency range. If increasing the coupling seems to have no broadening effect on the peak, the coupling should be reduced until an adjustment is found where the single peak flattens out, or the two humps begin to appear within the available frequency span.

To obtain the desired output characteristic from a bandpass coupler, it is essential that the excitation at the grid of the amplifier be substantially constant across the band for which operation is contemplated.

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

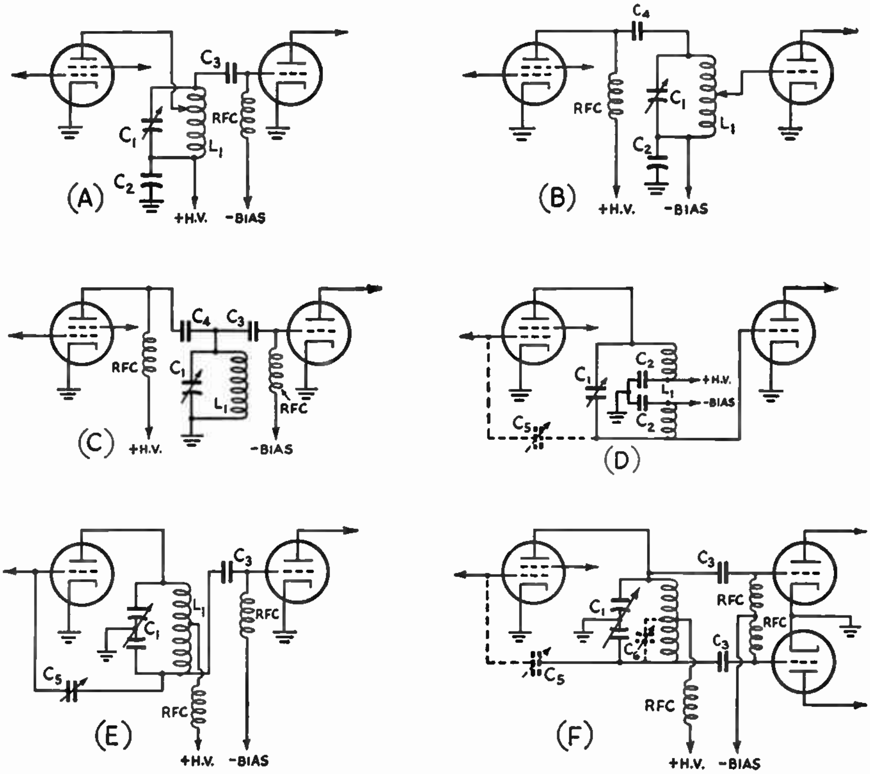

Fig. 6-17 — Examples of capacitive coupling. A — Series plate feed, parallel grid feed. B — Parallel plate feed, series grid feed. C — Parallel feed in both plate and grid. D — Series feed in both plate and grid. E — Balanced output to unbalanced input, series plate feed, parallel grid feed. F — Single 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.

balanced and unbalanced circuits, depending upon whether series or parallel power feed is desired, are shown in Fig. 6-17.

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 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. 6-17A 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 an alteration of coupling, however, will necessitate readjustment of neutralization if the tank is used for neutralizing the driver as suggested.

The circuit of Fig. 6-17E 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.

Capacitive-Coupling Adjustment

With capacitive coupling, an exact match between output and input impedances that differ appreciably cannot be made without tapping the coil (Figs. 6-17A and B). Tapping not only often presents a mechanical problem, but also is sometimes responsible for parasitic oscillation. When the output impedance of the driver is lower than the input impedance of the driven stage, the plate of the driver should be tapped down on the coil. On the other hand, if the input impedance of the driven tube is the lower, the grid should be tapped down.

In many cases, satisfactory coupling can be obtained without tapping the coil, particularly if the impedances to be coupled do not differ too greatly, or if the driver has power to spare. Within limits, the loading of the driver can be changed by a change in the capacitance of the coupling condenser, a smaller condenser reducing the coupling. Whether a tap is necessary and where it should be placed can be determined as discussed under "Adjustment of R.F. Amplifiers," this chapter.

Coupling Condensers

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

Amplifier Design Considerations

● PLATE-CIRCUIT VALUES

Tank-Circuit Q

The primary function of the plate tank circuit in a Class C amplifier is that of smoothing out the pulses fed to it by the tube. As discussed in the chapter dealing with electrical laws and circuits, it is also an essential part of certain systems used to couple the tube output to a useful load circuit. For both of these objectives, the Q of the tank circuit when shunted by the load should have a certain minimum value. With the condition under which Class C r.f. power amplifiers are usually operated in amateur transmitters, a minimum Q of 12 is recommended. For a fixed value of Q, the tank-circuit L/C ratio varies in proportion to the ratio of the d.c. plate voltage to plate current with the amplifier loaded. The chart of Fig. 6-18 shows required values of tank capacitance for a Q of 12 for a wide range of plate-voltage/plate-current ratios for each of the lower-frequency amateur bands. The values apply to the type of plate tank circuits shown in Fig. 6-19A only. Because the tube is connected across only half of the tank in the remainder of the circuits shown in Fig. 6-19, 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. 6-18 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. 6-19 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. 6-19, if power is reduced temporarily while tuning up without load.

In the circuits of Fig. 6-19C, 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

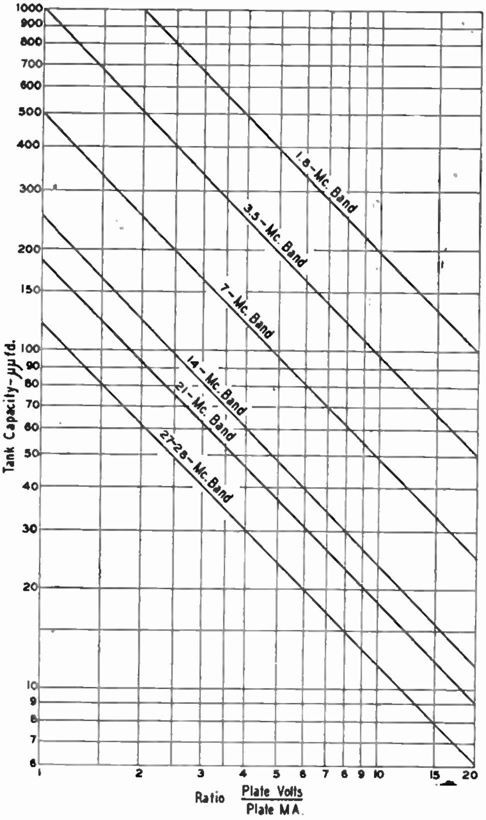

Fig. 6-18 — Chart showing minimum plate tank capacitances recommended with various ratios of plate voltage to plate current, for the six low-frequency amateur bands. In the circuits F, G and H of Fig. 6-19, the values shown by the graph may be divided by four. In circuits C, D, E, I, J and K, the capacitance 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.

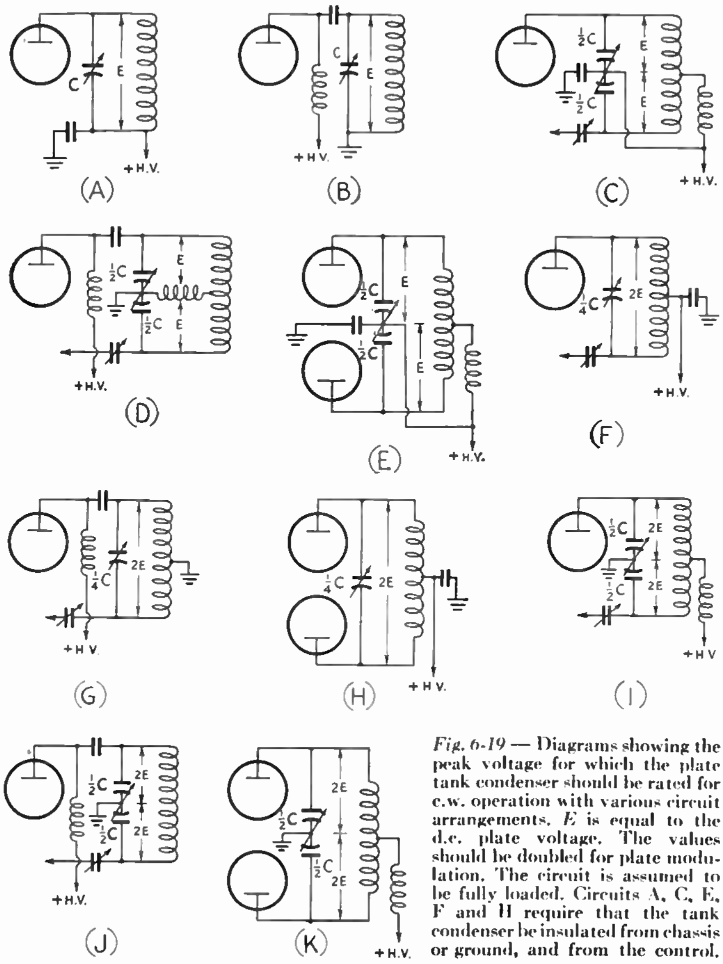

Fig. 6-19 — Diagrams showing the peak voltage for which the plate tank condenser should be rated for c.w. operation with various circuit arrangements. *E* is equal to the d.c. plate voltage. The values should be doubled for plate modulation. The circuit is assumed to be fully loaded. Circuits A, C, E, F and H require that the tank condenser be insulated from chassis or ground, and from the control.

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. 6-18 will be greater than 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 may 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. Disk-type ceramic condensers with capacitances between 0.001 and 0.01 μfd. are recommended where the plate voltage does not exceed 400 or 500 volts. Mica condensers of similar capacitance should be used at higher voltages. The voltage rating should be 25 to 50 per cent higher than the plate voltage for c.w. and double this value for a stage that is to be plate modulated.

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 provided it can be fitted with a protective guard.

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 about 125 ma. In the circuit of Fig. 6-19D, 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

As with the plate tank circuit, the grid tank-circuit Q should have a certain minimum value to assure adequate coupling. The amount of capacitance necessary for a given grid-tank Q depends upon the average grid impedance shunting the tank circuit. As the equation given later in the section on r.f. power-amplifier-tube operating factors indicates, this impedance under usual Class C operating conditions is a function of driving power and d.c. grid current. Fig. 6-20 shows the various values of grid tank capacitance required for a Q of 12 for different ratios of driving power to the square of the d.c. grid current. The value of C thus determined is for the single-tube circuit of Fig. 6-21A. This value should be modified as indicated in other sections of the same figure for tubes in push-pull.

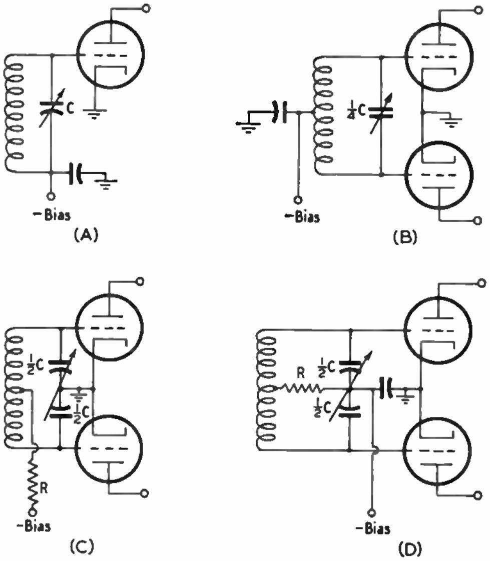

Fig. 6-21 — **Grid tank circuits. The chart of Fig. 6-20 gives the proper value for *C*. The tank condenser should have a voltage rating approximately equal to the operating bias voltage plus 20 per cent of the plate voltage for circuit A, twice this value for circuit B and each section of the condenser in circuit D. 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 minimum, which must be considered as part of the grid leak.**

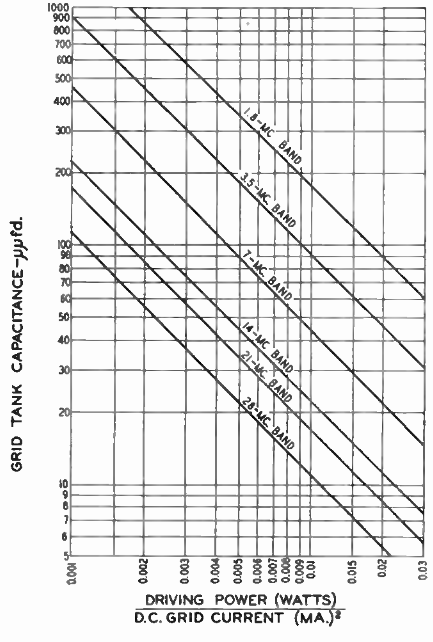

Fig. 6-20 — **Chart showing required grid tank capacitance for a *Q* of 12 for various ratios of driving power to d.c. grid current squared. These values should be modified as indicated in Fig. 6-21 for circuits other than 6-21A.**

If it is found that a Q as high as 12 is not needed to provide adequate coupling, the tuning of the grid circuit can be broadened by increasing the L/C ratio. This makes it necessary to retune the grid circuit less frequently in covering a given band.

Approximate tank-condenser voltage ratings are suggested under Fig. 6-21. 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. 6-21C 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.

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. Some transmitting-tube manufacturers list two sets of ratings, one for continuous commercial service and one for intermittent amateur (ICAS) service.

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 of excitation.

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, 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 power-supply chapter), 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. 6-22. 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. 6-22B. 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. 6-22C, 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. 6-22D, 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. 6-22E. The voltage across the regulator tube remains constant with or without grid current flowing. By making the voltage-regulator series resistor, R_4, 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. 6-22C. The use of VR tubes for this purpose is discussed more fully in the power-supply chapter.

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. 6-22F, 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

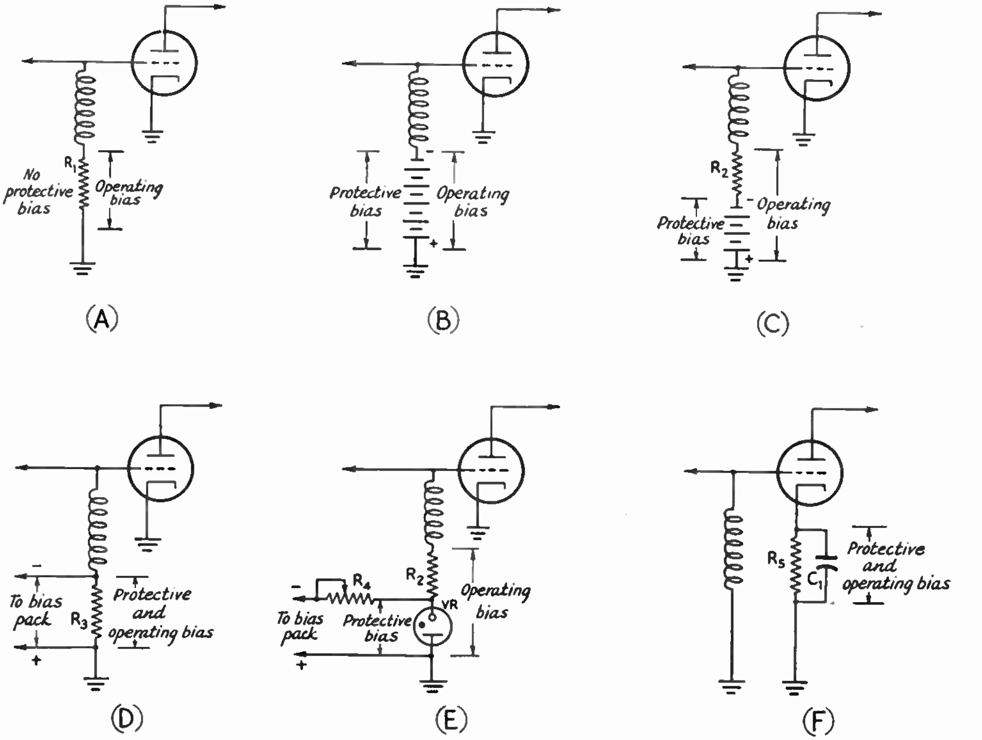

Fig. 6-22 — Various systems for obtaining protective and operating bias for r.f. amplifiers. A — Grid-leak. B — Battery. C — Combination battery and grid leak. D — Grid leak and adjusted-voltage bias pack. E — Combination grid leak and voltage-regulated pack. F — Cathode bias.

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. 6-22A, 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. 6-22D).

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.

Grid Impedance

For the conditions under which Class C amplifiers normally are operated, the impedance presented by the grid may be calculated to a close approximation by:

Input impedance (ohms)

$$= \frac{\textit{driving power} \text{ (watts)}}{\textit{d c. grid current} \text{ (ma.)}^2} \times 622 \times 10^3.$$

Both the driving power and the grid current are given in tube-operating data. The single-tube value thus obtained should be doubled for tubes in push-pull or halved for two tubes in parallel. These values will, of course, change if the operating conditions vary from those specified.

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 the relative driving power delivered to the input circuit. Fig. 6-23 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.

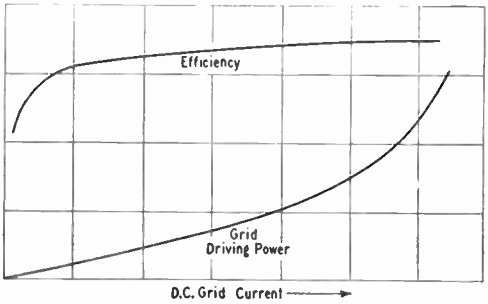

Fig. 6-23 — Curve showing relation between driving power and plate-circuit efficiency of an r.f. power-amplifier stage.

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.

Optimum Plate Load Resistance

The approximate optimum load resistance for a tube working under usual Class C operating conditions may be calculated by using the following equation:

$$\textit{Optimum plate load} \text{ (ohms)} = 570 \times \frac{\textit{d.c. plate voltage}}{\textit{d.c. plate current} \text{ (ma.)}}.$$

For a push-pull amplifier the value thus calculated for a single tube should be doubled, and halved for two tubes in parallel.

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. 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.

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. 6-24 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. 6-25 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. 6-24. 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 on measuring equipment.) 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 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.

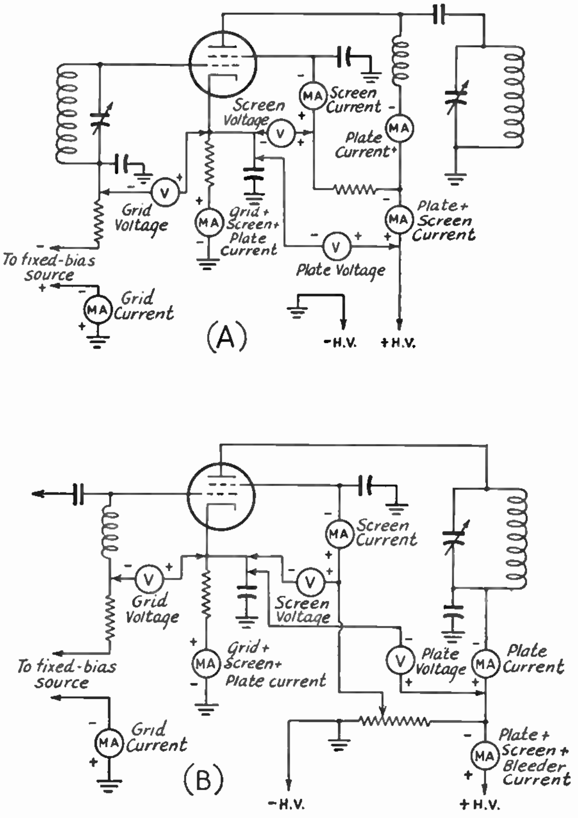

Fig. 6-24 — Diagrams showing placement of voltmeter and milliammeter to obtain desired measurements. A — Series grid feed, parallel plate feed and series screen voltage-dropping resistor. B — Parallel grid feed, series plate feed and screen voltage divider.

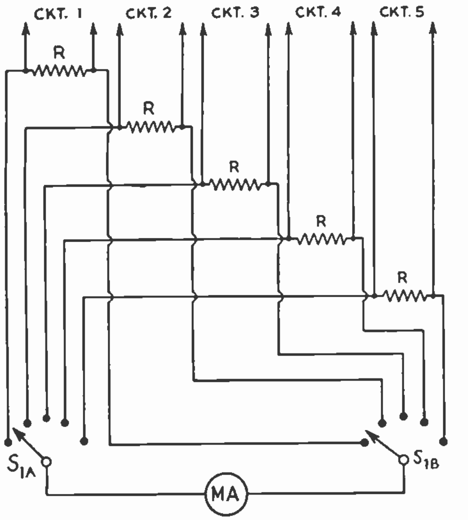

Fig. 6-25 — Method of switching a single milliammeter. The resistors, *R*, should be 10 to 20 times the internal resistance of the meter; 47 ohms will usually be satisfactory. S_1 is a 2-section rotary switch. Its insulation should be ceramic for high voltages, and an insulating coupling should always be used between shaft and control knob.

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. 6-10. 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 adequate coupling cannot be obtained, that is, if the grid current is below normal and the driver plate current is still below the rated maximum operating value, the earlier sections on coupling considerations should be reviewed.

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 readjustment 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, 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, 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.

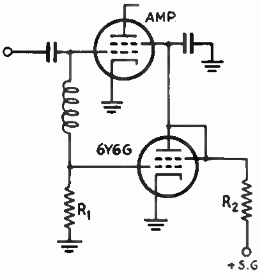

Fig. 6-26 — Screen protective circuit for screen-grid amplifier as an alternative to the use of fixed bias. R_1 is the normal grid leak for the amplifier and R_2 the recommended screen voltage-dropping resistor.

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 maintain it at rated value.

A screen-grid tube should never be operated at full screen voltage without plate voltage and full load. The screen may be damaged 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. 6-26. 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. This system usually is not satisfactory if the stage is to be screen-and-plate modulated.

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. 6-27. At A a thermoammeter, A, and a noninductive (ordinary 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 noninductive 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:

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

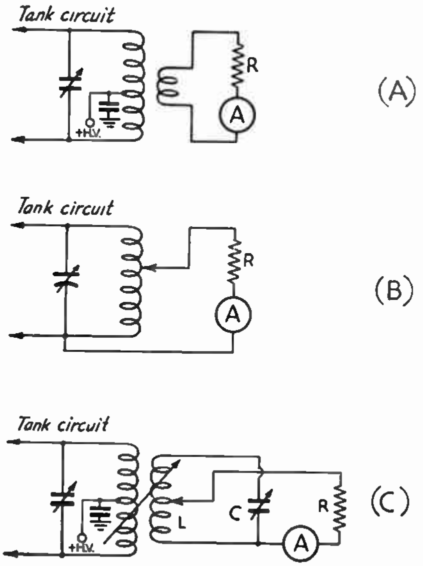

Fig. 6-27 — "Dummy-antenna" circuits for checking power output and making adjustments under load without applying power to the actual antenna.

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. 6-27B 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. 6-27B is used.

If the dummy load resistance chosen is the same as the impedance of the antenna system to be used (assuming a resonant system or one in which the line and the antenna have been matched), the coupling adjustments made for the dummy should hold for the antenna also.

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 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. 6-28. 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.

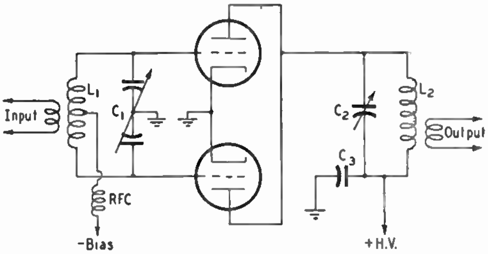

Fig. 6-28 — Circuit of a push-push frequency multiplier for even harmonics. The grid tank circuit, L_1C_1, is tuned to the frequency of the preceding driving stage, while the plate tank circuit, L_2C_2, is tuned to an even multiple of that frequency, usually the second harmonic. C_3 is the plate by-pass capacitor, usually a 0.001-μfd. mica condenser, while *RFC* is a 2.5-mh. r.f. choke.

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 as to frequency and type or mode.

● 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. 6-29A. 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 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. A 150-watt size is about right for a medium-power transmitter.

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 measurements chapter) tuned to the frequency of the parasitic and held close to the plate lead of the tube.

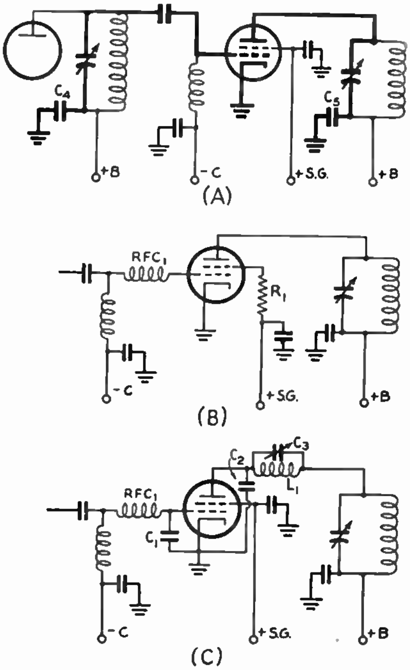

Fig. 6-29— A — V.h.f. parasitic circuit hidden in high-frequency amplifier. B — One method of suppression with tetrodes. C — Preferred method. Approximate values: RFC_1 — 15 turns No. 22, ¼-inch diam., close-wound; C_1 — 12-μμfd. ceramic; C_2 — 15-μμfd. tubular; C_3 — 100-μμfd. midget variable; L_1 — 3 turns No. 14, ½-inch diam., ½ inch long.

Remedies

At the outset, an amplifier should be laid out so that the heavy leads shown in Fig. 6-29A 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.

It is often helpful to provide returns independent of the chassis by mounting the tank condensers on a wide aluminum strip as a subassembly with the tube socket also mounted on the strip.

The paths through the by-pass condensers (C_4 and C_5, Fig. 6-29A) 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. The use of 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 grid and plate 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. 6-29B. 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. 6-29C 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 measures that are necessary to take toward 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. 6-29C. The adjustment of such wavetraps in a push-pull amplifier will have a

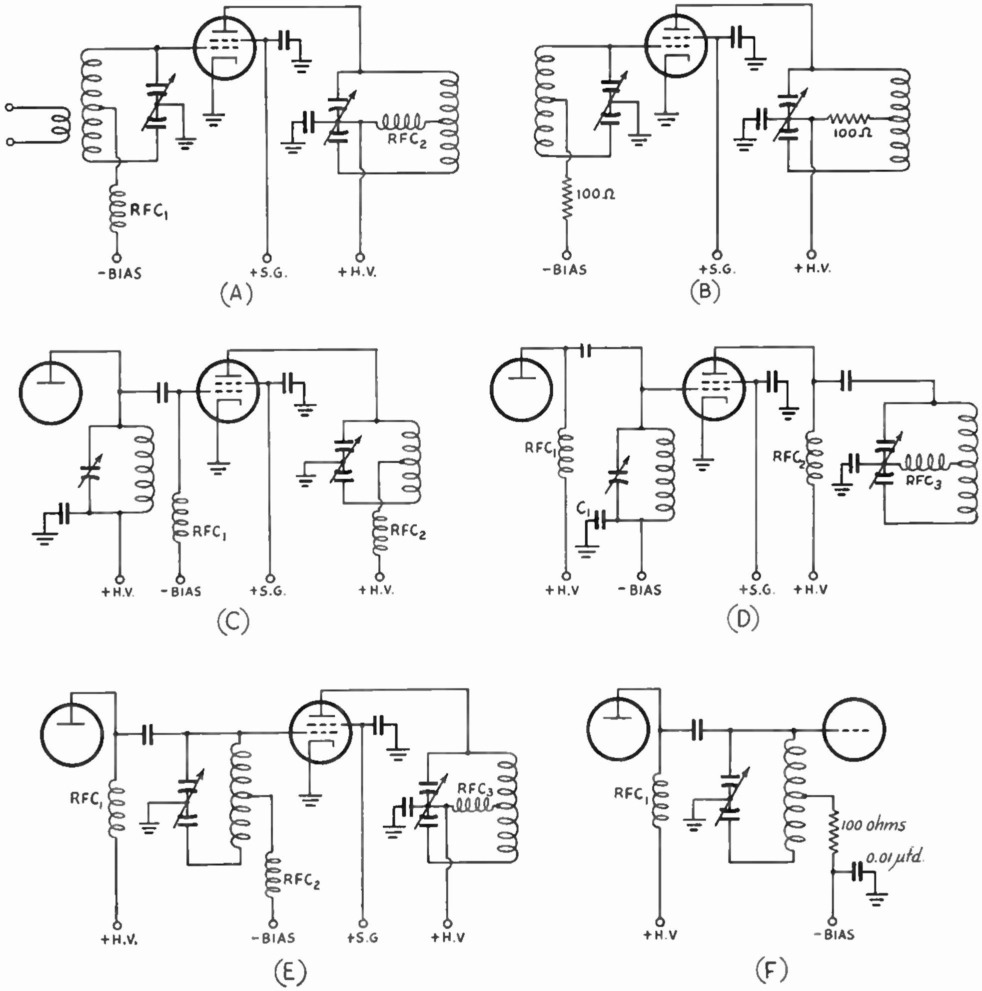

Fig. 6-30 — A, C and E show circuit arrangements to be avoided in eliminating low-frequency parasitic oscillation. The circuits of B, D and F are the recommended alternatives, which apply equally well to push-pull.

marked effect on the balance, however, and balance and suppression adjustments cannot be made independently.

A sensitive grid-dip meter of the type described in the measurements chapter is often helpful in the locating of resonances responsible for parasitic oscillation as well as those which act to reinforce v.h.f. harmonics. Often the elimination of one will result in elimination of the other. 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. However, 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 not permitted to exist.

Circuits To Be Avoided

Fig. 6-30 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. 6-30C, 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 applied to 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 reads minimum.

● NEUTRALIZING INDICATORS

Fig. 6-31 shows the diagram of a sensitive neutralizing indicator. By referring to the measurements chapter, 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.

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.

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

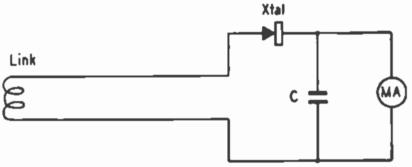

Fig. 6-31 — Circuit of sensitive neutralizing indicator. *Xtal* is a 1N34 crystal detector, *MA* a 0-1 direct-current milliammeter and *C* a 0.001-μfd. mica by-pass condenser.

through resonance. This dip in grid current reduces as neutralization is approached until at exact neutralization all change in grid current should disappear.

EXTERNAL COUPLING

Before making any neutralizing adjustment, the circuit should be checked for external coupling. This can be done by removing the tube from the socket and disconnecting the neutralizing condenser. If the plate terminal is at the socket, it should be disconnected also. With the driver in operation, the plate tank circuit should be checked for r.f. with one of the r.f. indicators previously suggested.

The input and output coils should be so placed in 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, if the coils are close together, it may be necessary to introduce shielding between the input and output circuits.

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 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, it should always be possible to find a point of minimum indication with an increase on either side.

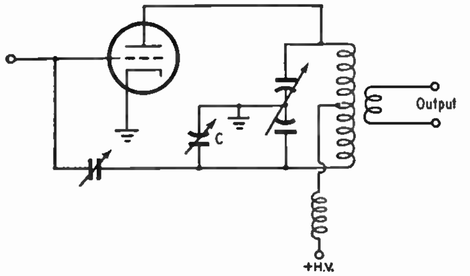

Fig. 6-32 — In this neutralizing circuit, *C*, which should have the same capacitance as the output capacitance of the tube, has been added to compensate for the tube capacitance across the upper half of the circuit.

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. 6-32 should be added and adjusted to match the output capacitance of the tube.

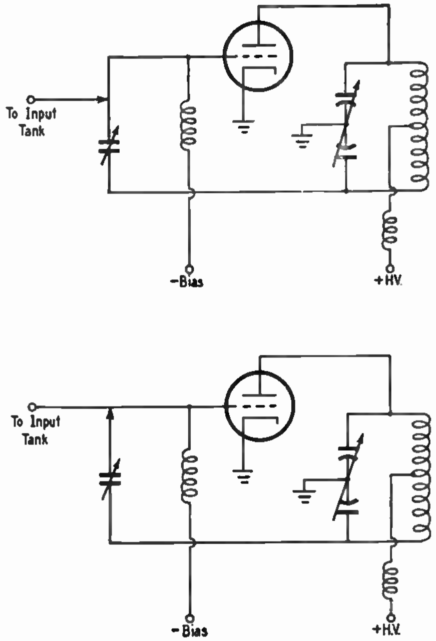

Fig. 6-33 — If an amplifier fails to remain neutralized on all bands, the condition usually can be remedied by tapping the input-tank lead along the neutralizing-condenser lead (or vice versa), adjusting the position until the amplifier neutralizes at the highest frequency at the same setting of the neutralizing condenser as at the lowest frequency. The same adjustment should be made to both sides of a push-pull circuit.

In an amplifier which is to be used on several bands, it should be first neutralized when 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. 6-33 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.

It will seldom be possible to reduce the indication to zero with a sensitive indicator unless excitation is reduced below the grid-current point.

Adjustment of Inductive Neutralizing

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

The inductive neutralizing system of Fig. 6-13C is adjusted by adjusting the value of either the inductance or the capacitance for minimum feed-through to the plate circuit.

A Single-Control Low-Power Transmitter

Figs. 6-34 through 6-40 show the circuit and constructional details of a 40-watt two-stage transmitter that requires the adjustment of only one tuning control. The crystal oscillator uses a modified Pierce circuit. The use of bandpass couplers in the output circuit of this stage makes it unnecessary to retune when changing frequency and at the same time provides inductive coupling as a measure toward reducing v.h.f. harmonics. The coupling between the two circuits is adjusted to give the desired broadband response and then fixed in that position. It is possible to arrive at an adjustment where the amplifier grid excitation is substantially constant over any given band and drops off quite sharply outside the band edges. (See Fig. 6-16.)

The output stage is a conventional 807 amplifier normally working straight through on the output frequency of the oscillator, except for 28 Mc., although it will double frequency to any of the lower-frequency bands. RFC_3 and R_6 are parasitic suppressors. The amplifier grid leak, R_5, is connected in series with the grid tank circuit, since the coupler provides an opportunity to avoid parallel grid feed. RFC_4 and C_{12}, RFC_5 and C_{13} are v.h.f. harmonic filters.

The unit is designed to operate from a single power supply delivering 300 to 450 volts. To avoid the need for fixed bias on the output stage, both stages are keyed simultaneously in the common cathode lead. The octal socket used as a crystal mounting also provides a means of feeding a VFO into the unit. Connections are shown in Fig. 6-39.

Construction

The transmitter is built in a standard 5 × 9 × 6-inch steel utility box. Most of the parts are mounted on an aluminum plate cut to fit the inside of the box and supported from its sides by ½-inch angle brackets as shown in the bottom view of the unit, Fig. 6-36. The plate is mounted 3⅝ inches above the bottom of the box. Two ventilating holes are cut through the plate near the front of the box, and additional vents are punched through the top and bottom covers of the box. These holes permit air to circulate through the entire box, yet do not reduce the effectiveness of the shielding.

The sockets for the 6AG7 and for the plug-in bandpass coupler are mounted in line, 1¼ inches from the rear of the aluminum plate. The socket for the 807 is mounted in a Millen bracket assembly (80007) trimmed down to fit below in a horizontal position. It is placed so that the grid terminal is 3¾ inches from the rear of the box, allowing adequate space for mounting the small parts in the oscillator circuit, yet retaining the desired short r.f. leads.

An octal socket used to hold the crystal and to connect a VFO, an octal plug for power input connections, and a coaxial output connector are mounted at the rear, centered 1¼ inches above the bottom edge. The key jack and a panel light are mounted on the front, spaced 1⅝ inches above the bottom edge.

The top view of the transmitter, Fig. 6-34, shows the arrangement of the plate tank circuit of the 807 stage. A six-prong ceramic socket for the plug-in plate coils is supported above the deck by ¾-inch ceramic stand-off insulators (National GS-10) 4⅞ inches behind the front of the box. The tuning condenser is mounted on ceramic button-type insulators (National XS-6) immediately in front of the coil socket. The rotor shaft of this condenser

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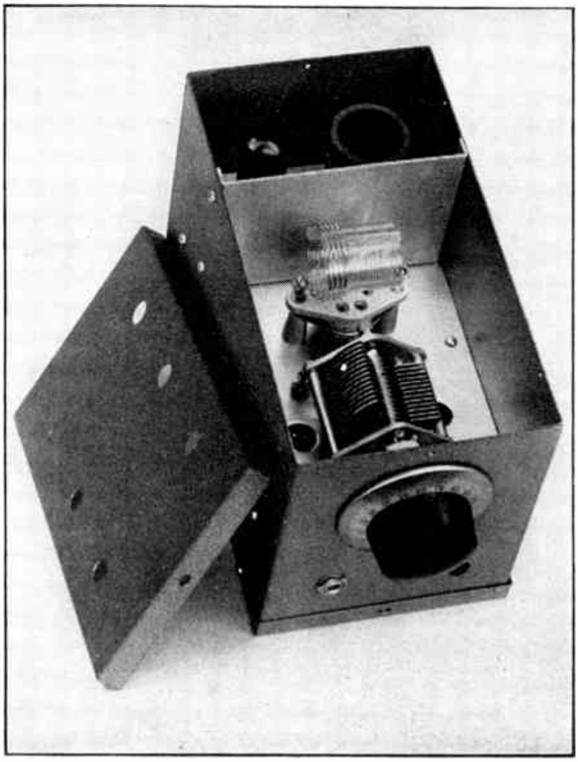

Fig. 6-34 — Front view of the transmitter with cover removed. The tank circuit for the 807 amplifier occupies the front compartment, with the 6AG7 oscillator and the plug-in bandpass coupler at the rear. Ventilation for the tubes is obtained through holes punched in the top, bottom, and the interior mounting plate which supports the various components.

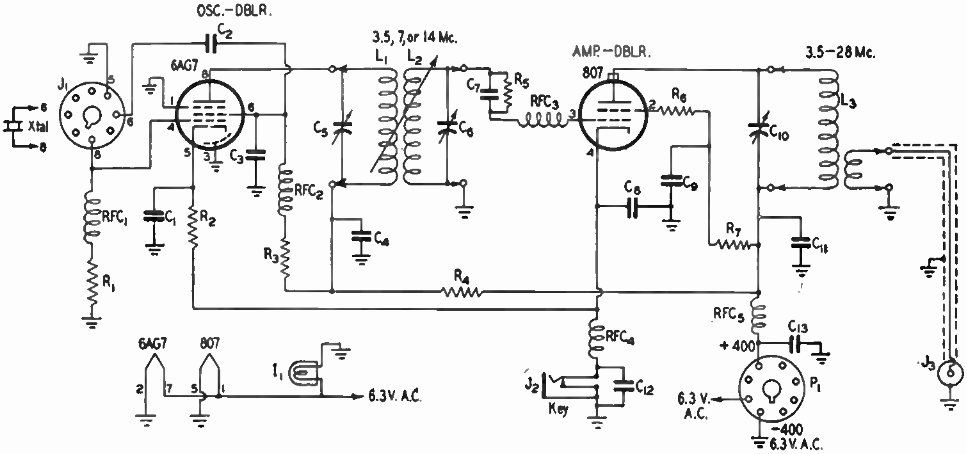

Fig. 6-35 — Circuit diagram of a two-stage four-band transmitter utilizing bandpass coupling and including TVI-reducing filters.

C_1, C_8, C_9 — 0.01-μfd. disc ceramic.
C_2 — 0.005-μfd. disc ceramic.
C_3 — 25-μμfd. mica.
C_4, C_{12}, C_{13} — 0.001-μfd. disc ceramic.
C_5, C_6 — 3–30 μμfd. air-dielectric trimmers (Phillips).
C_7 — 100-μμfd. mica.
C_{10} — 300-μμfd. transmitting variable (National TMS-300).
C_{11} — 0.001-μfd. mica, 1200 v. d.c. working.
R_1 — 47,000 ohms, ½ watt.
R_2 — 330 ohms, 1 watt.
R_3 — 47,000 ohms, 1 watt.
R_4 — 10,000 ohms, 5 watts, wire-wound.
R_5 — 22,000 ohms, 1 watt.
R_6 — 47 ohms, ½-watt carbon.
R_7 — 20,000 ohms, 5 watts, wire-wound.
L_1 — Primary, bandpass coupler.
  3.5 Mc. — 40 turns No. 30 d.s.c., close-wound, 1½-inch diam. form.
  7 Mc. — 16 turns No. 26, d.s.c., close-wound, 1½-inch diam. form.
  14 Mc. — 9 turns No. 20 d.s.c., close-wound, 1½-inch diam. form.
L_2 — Secondary, bandpass coupler. Wound on same form as L_1, spaced as indicated.
  3.5 Mc. — 35 turns No. 30 d.s.c., close-wound, ¼-inch separation from L_1.
  7 Mc. — 15 turns No. 26 d.s.c., close-wound, 9/16-inch separation from L_1.
  14 Mc. — 9 turns No. 20 d.s.c., close-wound, ½-inch separation from L_1.
L_3 — Plate coil for 807 stage. (All are National AR-17 series).
  3.5 Mc. — AR-17-40E. (28 turns No. 18, 1 9/16 inches long, 1¼-inch diam.)
  7 Mc. — AR-17-20E. (14 turns No. 16, 1¼ inches long, 1¼-inch diam.)
  14 Mc. — AR-17-10E. (8 turns No. 16, 1⅝ inches long, 1¼-inch diam.)
  28 Mc. — AR-17-6E. (4 turns No. 12, 2 inches long, ⅞-inch diam.)
I_1 — 6.3-volt pilot lamp.
J_1 — Octal socket, ceramic.
J_2 — Closed-circuit jack.
J_3 — Coaxial connector, female.
P_1 — Octal plug, panel mounting.
RFC_1, RFC_2 — 2.5-mh. r.f. choke (National R-100-S).
RFC_3 — 1.8-μh. r.f. choke (Ohmite Z-144).
RFC_4, RFC_5 — 7-μh. r.f. choke (Ohmite Z-50).

must be insulated from the front panel because it carries the full plate-supply voltage. The shaft is 1½ inches above the aluminum plate when mounted as described, and passes through the front of the box 2 inches below the top. The two leads that connect the condenser to the tube and to the plate by-pass condenser pass through the mounting plate in polystyrene through-bushings (National TPB).

An aluminum partition 3⅜ inches high divides the top portion of the box into two compartments, providing shielding between the bandpass coupler and the plate coil of the 807. These two coils are mounted at right angles to each other as additional insurance against feed-back.

The coaxial output link runs from the prongs of the coil socket through a ¼-inch hole in the plate to the output connector on the rear of the box. Both ends of the shield braid of this link are grounded to the chassis.

The components used to filter the d.c. leads (RFC_4, RFC_5, C_{12}, and C_{13}) are mounted as close as possible to the points where the leads pass through the shield enclosure, using very short leads from the condensers to ground. Parasitic-suppressing choke RFC_3 is mounted right at the grid terminal of the 807 socket, and R_6, which also has a part in eliminating parasitics, is mounted between the screen-grid terminal and a small tie-point bolted to the mounting bracket. Screen by-pass condenser C_9 is connected from this tie-point to the cathode pin on the tube socket. Plate by-pass condenser C_{11} is placed behind the 807, between it and the mounting plate which serves as ground. The lead from the "high" side of this condenser to the plate tank circuit passes through a bushing.

All heater and d.c. wiring is made with shielded wire, with the braid grounded at each end. The screen dropping resistor R_7, and R_4, which reduces the supply voltage to the proper level for the oscillator, are mounted on tie-points near the octal power plug.

The circuit diagram of a power supply for this transmitter is shown in Fig. 6–40. It is conventional with condenser-input filter. A separate filament transformer is provided so that the plate supply may be turned off independently.

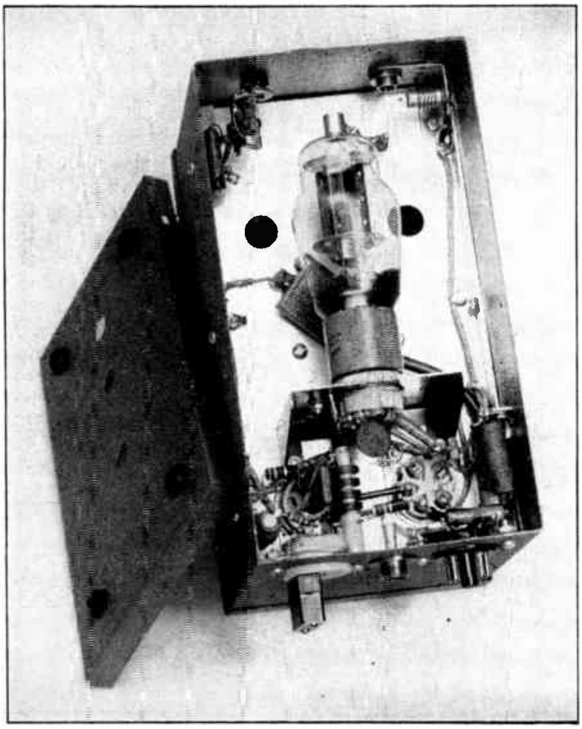

Fig. 6-36 — Bottom view of the transmitter. The 807 socket is mounted in a cut-down commercial bracket, with the sockets for the 6AG7 and the bandpass coupler spaced below and to either side of it. Arranged along the rear of the box are the crystal socket, the output jack, and the power plug.

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The diagram of a suitable power supply for this unit appears in Fig. 6-43. A separate filament transformer is suggested so that the high voltage may be turned off independently of the filaments.

Bandpass Couplers

Three couplers are needed to use the transmitter in four amateur bands. One coupler is designed to provide excitation across the entire 3.5–4-Mc. band, another for the 7–7.3-Mc. band, and the third from 14 to 14.9 Mc. This latter range is considerably in excess of what would be required for coverage of the 14-Mc. band alone. The extension at the high-frequency end of the range is necessary if the transmitter is to operate in the 28-Mc. band, because for output in this range, the 807 stage must be operated as a doubler from the 14-Mc. excitation supplied to its grid circuit.

In crystal-controlled operation, 3.5-Mc. fundamental crystals may be used for output in the 3.5- and 7-Mc. bands, and 7-Mc. crystals for output in the 7-, 14-, and 28-Mc. bands. In instances where a VFO is used to replace the crystal, the 6AG7 stage should be used as a frequency doubler to eliminate the possibility of oscillation.

The photograph of Fig. 6-37 and the sketch of Fig. 6-38 show how the bandpass couplers are constructed and wired. The Phillips trimmers are especially well adapted for this use, since they are readily mounted by inserting and soldering their spike terminals, along with the coil ends, into the pins of the National Type XR-5 coil forms. It is highly important that the windings be made as close as possible to the dimensions given under Fig. 6-35. It is perhaps advisable to not make the turns too snug on the form so that the distance between the coils can be given a final adjustment should this be found necessary.

The adjustment of the bandpass couplers can be checked by measuring the amplifier biasing voltage as the oscillator is tuned across the band. This can be done by connecting a high-resistance voltmeter between the 807 grid and ground, with a 2.5-mh. r.f. choke in series with the meter lead that is connected to the grid.

The checking should be done with the plate- and screen-voltage line to the 807 disconnected. Choose a crystal as close to the center of the band as possible and adjust C_5 and C_6 for maximum 807 grid voltage. The two adjustments will not be entirely independent, because of the coupling, and some juggling back and forth may be required before the setting for maximum reading is attained. Now, without further adjustment of the

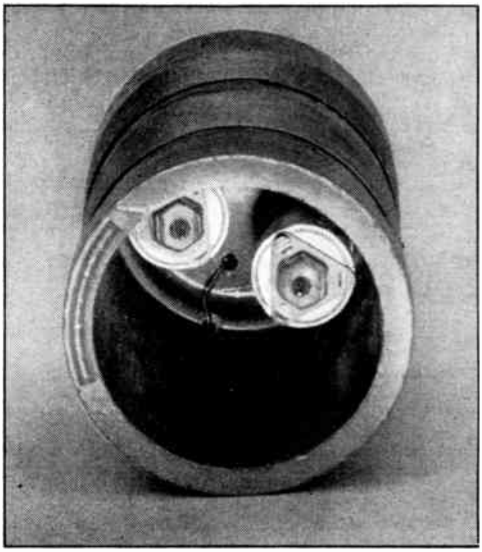

Fig. 6-37 — One of the bandpass couplers. The two trimmer condensers are mounted inside of the coil form, with connections made as shown in Fig. 6-41.

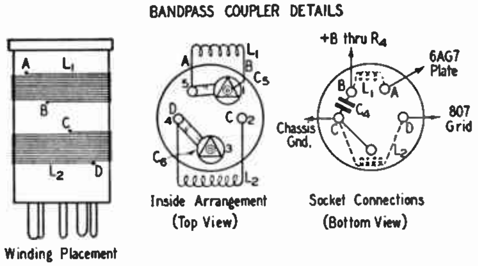

Fig. 6-38 — Details of the bandpass couplers. The trimmer condensers are soldered inside of the coil form, as described in the text, making a simple, compact plug-in assembly that needs adjustment only once.

coupler, plug in other crystals for the same band. If it is found that the grid voltage falls off considerably with crystals whose frequencies lie near the edges of the band, the windings should be moved slightly closer together and the check across the band made again. If it is found that the voltage is high near both ends of the band, but low in the middle, the coupling should be loosened. When the voltage is considerably higher at one end than the other, this can usually be corrected by trial readjustments of C_5 and C_6 in small amounts. For crystal control, it is necessary to carry the adjustment only to the point where adequate excitation (at least 45 volts bias with the amplifier running and loaded) is obtained with each of the available crystals. If a VFO is used, its output frequency should be one frequency band lower than the band of the coupler and the adjustments will have to be more exact if uniform excitation across the band is desired. Some means should be provided for adjusting the output of the VFO, since excessive driving of the 6AG7 may have an effect on the shape of the excitation curve.

Once the couplers are adjusted properly, the windings should be cemented in place with coil dope, and the rotors of the trimmers should be locked in position with a drop of Duco cement.

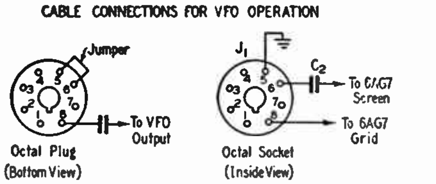

Fig. 6-39 — Method of substituting a VFO for the crystal. An octal plug, wired as shown, is inserted in the crystal socket. The jumper between Pins 5 and 6 serves to ground one side of C_2, thereby changing it from a coupling condenser to a screen by-pass condenser. Excitation from the VFO is applied to the grid of the 6AG7 through Pin 8 of the plug, which is connected to the center conductor of a short length of coaxial cable. The condenser shown at Pin 8 should be mounted inside the VFO, serving as a d.c. blocking condenser. Its size may be anything from 100 $\mu\mu$fd. to 0.001 μfd., with the smaller value being preferred.

Amplifier Adjustment

Reconnect the d.c. screen lead to the 807 stage, and plug a milliammeter capable of reading up to 200 ma. in the key jack where it will read the total current flowing in both stages. The 6AG7 plate current normally will run between 10 and 15 ma., so this should be subtracted from the meter reading to determine the current flowing in the 807. Plug the desired coil in the 807 plate circuit, and the correct crystal-coupler combination in the oscillator stage. Connect a 25-watt lamp bulb to the output terminal to serve as a dummy load while the 807 stage is tested.

Apply plate voltage and resonate the 807 tank circuit by tuning C_{10}. The off-resonance plate current will be very high, in the neighborhood of 200 ma., dipping to 100 ma. or less at resonance. If it is not possible to load the 807 stage so that the total current indication is 100 ma. or slightly over, disconnect the lamp from the output terminal and tap it across a few turns of the tank coil. This should be done with the power off, of course! By changing the number of turns across which the lamp is tapped and re-resonating the plate circuit, it should be possible to obtain full loading.

Check the keying characteristic by listening to the signal, or a harmonic of it, in the receiver with the gain turned down as far as possible and the antenna disconnected. With the circuit constants shown and active crystals, good keying should be obtained with both 3.5- and 7-Mc. crystals. If, however, the keying is sluggish, and it sounds as though the crystal doesn't start oscillating readily, the size of feed-back condenser C_3 should be changed in 25-$\mu\mu$fd. steps until good keying is obtained.

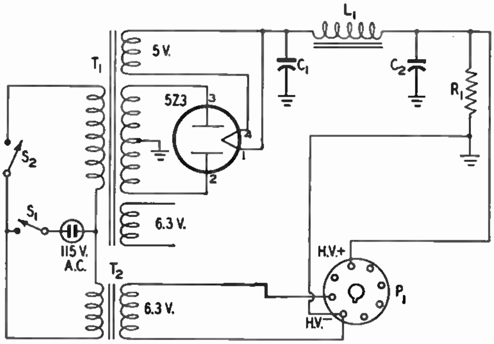

Fig. 6-40 — Diagram of a power supply for the single-control low-power transmitter.

C_1 — 2-μfd. 1000-volt oil-filled.
C_2 — 2-μfd. min. 1000-volt oil-filled.
R_1 — 15,000 ohms, 25 watts.
L_1 — 10 h. min., 130 ma. min.
P_1 — Octal female plug.
S_1, S_2 — 3-amp. toggle switch.
T_1 — Power transformer: 400 to 450 volts r.m.s. each side of center, 130 ma. min.; 5 volts, 3 amp. (6.3-volts, 1.5 amp. min. if used. See text.)
T_2 — Filament transformer: 6.3 volts, 1.5 amp. min.

An All-Band Bandpass Exciter

Figs. 6-41 through 6-46 show diagrams and constructional details of a 120-watt VFO-bandpass transmitter or exciter for a higher-power amplifier. An f.m. modulator and crystal calibrator, to be included if desired, are also described. Referring to the circuit diagram of Fig. 6-42, a 6AG7 series-tuned VFO, operating in the 2-Mc. range, doubles frequency to the 3.5-Mc. range and drives a second 6AG7 as a straight amplifier at this frequency. This amplifier then drives a series of 6SN7 frequency doublers (one triode section in each stage). The appropriate stages for any desired band are connected in by the bandswitch, S_1. The bandswitch also connects the grids of a parallel-connected 829-B output amplifier to the corresponding doubler output circuit. Bandpass couplers that provide, without retuning, essentially constant output over the bands in which they operate are used between all stages, instead of capacity coupling. Aside from the fact that the use of these couplers reduces the number of tuning controls to only 2 for a 7-stage transmitter, it also provides inductive coupling that discriminates against harmonics that may cause TVI. Since the bandpass circuits cannot be conveniently made to cover the wide frequency range including both the 28- and 27-Mc. bands, a separate coupler is used for each. A trimmer is connected across the input of each of the doubler stages to make the input capacitance equal to that of the 829-B so that the tuning of the couplers will not be disturbed when switching bands.

The output tank circuit of the 829-B amplifier is a combination "all-band" circuit that covers all bands without changing coils. At 3.5 and 7 Mc. $C_{54}L_{16}L_{17}$ act as a parallel-tuned circuit with the two sections of C_{54} connected across L_{17}. L_{16} may be considered as a jumper connection between the stator sections of the capacitor at frequencies below 7 Mc. However, the reactance of L_{16} becomes appreciable at 14 Mc. and above. At these frequencies the circuit becomes rather complex, consisting of the resultant of L_{16} and L_{17} partially in parallel, tuned by the resultant of the two sections of the tank capacitor in series. Two output-coupling links, both series-tuned by C_{59}, are terminated at one of the wafers on the bandswitch. L_{18} is the low-frequency link and L_{15} operates at 14 Mc. and above.

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The tubular condenser, C_{52}, and the wave-trap consisting of L_{14} and C_{53} represent measures taken to eliminate parasitic oscillation and v.h.f. harmonics in the output stage. In addition, all external power leads are provided with v.h.f. filtering and the meters are bypassed.

The f.m. modulator consists of a 6AK5 speech amplifier and a similar tube in the reactance modulator. The modulator is connected across the oscillator tank circuit.

The crystal calibrator is a simple 100-kc. oscillator giving 100-kc. points throughout the spectrum. Thus band-edge markers are provided for setting the VFO.

The transmitter requires two power supplies, one delivering 300 volts, 150 ma. and the other 600 to 750 volts, 200 ma. A source of fixed bias also is needed for the 829-B. This may consist of a 45-volt battery, as Fig. 6-42 indicates, or R_{25} may be omitted and the bias obtained from a VR-regulated supply. Fig. 6-46 shows the diagram of a suitable power unit, including provision for bias. In the transmitter, a 150-volt lead, regulated by a VR-150, is brought out from the 300-volt input terminal. From this is operated the plate and screen of the VFO, the screen of the 6AG7 buffer, the reactance modulator and the 100-kc. crystal calibrator. A 6.3-volt transformer that supplies all filaments is included in the transmitter.

Transmitter Construction

The chassis for the transmitter measures 3 by 10 by 17 inches and the whole unit is housed in a No. CA-304 Par-Metal cabinet. The rear view (Fig. 6-43) shows the meter shields mounted on the front panel to the left and right of the dial lamps and the output-link tuning condenser at the lower left-hand corner of the panel. The 829-B tube socket is sub-

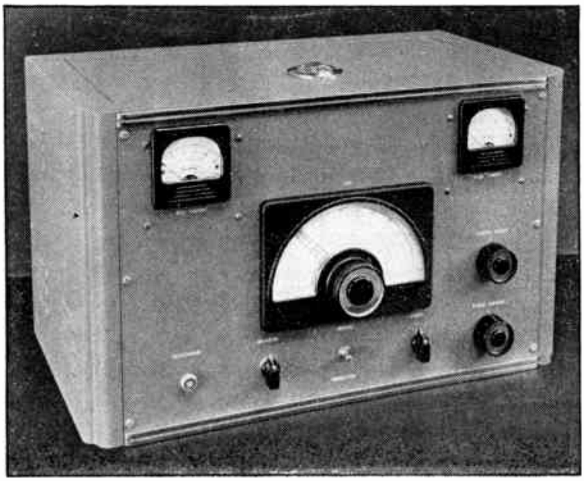

Fig. 6-41 — **Operating controls have been cut to a minimum in this bandswitching transmitter. Only the VFO dial need be tuned for coverage of a large portion of any one band. The grid and plate meters are to the left and right of the main tuning dial. The microphone jack, gain control, crystal-modulator switch, bandswitch and the amplifier control knob are in line across the bottom of the panel. The output-link tuning control is just below the plate meter.**

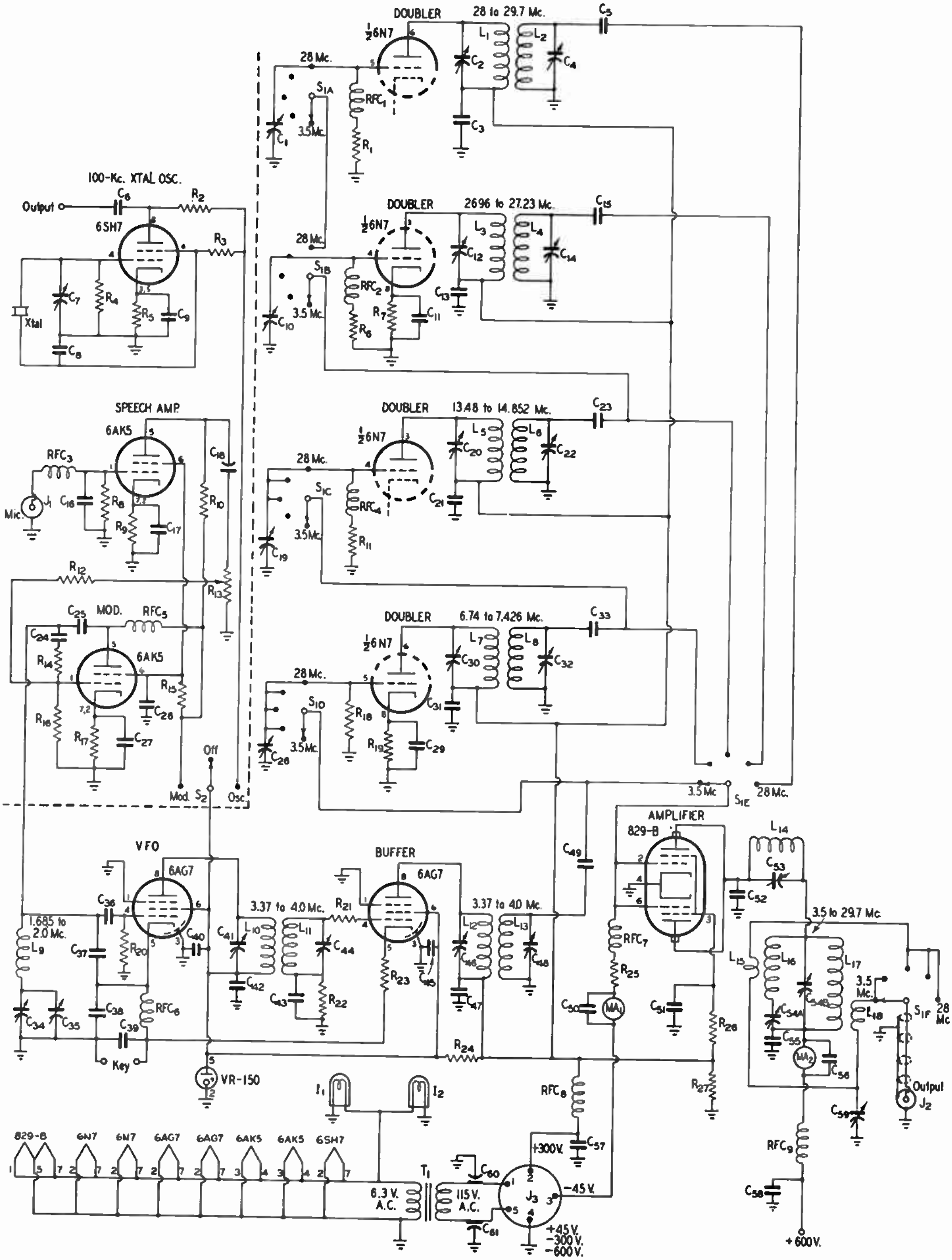

Fig. 6-42 — Circuit diagram of the bandpass transmitter.

C_1, C_2, C_4, C_{10}, C_{12}, C_{14}, C_{19}, C_{20}, C_{22}, C_{28}, C_{30}, C_{32}, C_{41}, C_{44}, C_{46}, C_{48} — 30-μμfd. ceramic trimmer (National M30).

C_3, C_{11}, C_{13}, C_{21}, C_{26}, C_{29}, C_{31}, C_{39}, C_{40}, C_{42}, C_{43}, C_{45}, C_{47}, C_{50}, C_{56} — 0.01-μfd. disc-type ceramic (Sprague 36C1).

C_5, C_{15}, C_{24}, C_{25} — 47-μμfd. mica.

C_6 — 22-μμfd. mica.

C_7 — 50-μμfd. variable (Millen 26050).

C_8 — 150-μμfd. mica.

C_9 — 0.0022-μfd. mica.

C_{16}, C_{23}, C_{33}, C_{36}, C_{49} — 100-μμfd. mica.

C_{17} — 10-μfd. 25-volt electrolytic.

C_{18} — 0.01-μfd. 400-volt paper.

C_{27} — 0.025-μfd. 400-volt paper.

C_{34} — 50-μμfd. variable (Millen 19050).

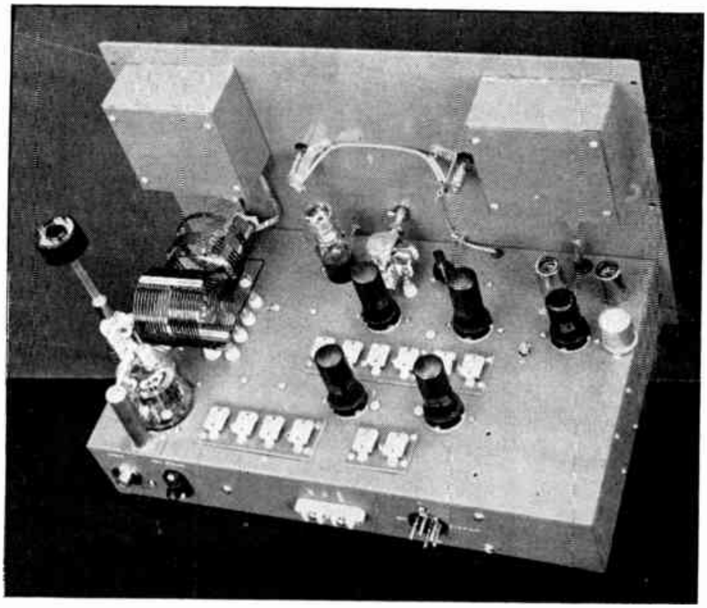

Fig. 6-43 — A rear view of the bandpass transmitter. Rectangular holes, cut in the chassis, provide clearance for the coupler coils. The coupler capacitors are readily accessible for adjustment.

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mounted at the left end of the chassis in between the homemade tubular condenser (Fig. 6-44) and the amplifier plate coils. A stand-off insulator supports the plate trap to the left of the 829-B. Feed-through insulators to the right of the plate coils allow connections to the output links and the bandswitch underneath. L_{15} is the 3-turn winding located closest to the panel and L_{18} is the coil in front of the 829-B (right).

To the right of the amplifier components, in the line nearest the panel, are the VR-150, C_{34}, the control for C_{35}, and the modulator tubes. The second line of parts starts at the left with the 7- and 14-Mc. doubler tube and continues to the right with the VFO tube, the 6SH7, and the 100-kc. crystal. To the rear of the first two tubes are the 14- and 7-Mc. couplers and the coupler in the output of the VFO. From left to right are C_{22}, C_{20}, C_{32}, C_{30}, C_{41} and C_{44}. The two tubes to the rear are the 10- and 11-meter doubler to the left and the 6AG7 buffer to the right. Behind these tubes are the 10- and 11-meter couplers and the coupler in the output of the 6AG7. Left to right are C_{14}, C_{12}, C_4, C_2, C_{48} and C_{46}.

The bottom view of the transmitter (Fig. 6-45) shows the amplifier tank condenser and the plate by-pass capacitor, C_{55}, lined up to the right in front of the 829-B tube socket. The tank capacitor, C_{54}, is insulated from ground (for d.c.) by means of National XP-6 polystyrene buttons and an insulated shaft coupling protects the operator from accidental contact with the "hot" control shaft. Screen-grid resistors, R_{26} and R_{27}, are mounted directly on the tube socket and the 829-B grid r.f. choke is located on the rear wall.

Aluminum brackets support the bandswitch at the right center of the chassis. The rear wafer of this switch accommodates the wiring for the 27- and 28-Mc. doubler tube, the center section takes care of the 14-Mc. 6N7 output circuit and

C_{35} — 100-μμfd. variable (Millen 20100).
C_{37}, C_{38} — 670-μμfd. silver mica.
C_{51} — 0.005-μfd. ceramic (Sprague 29C1).
C_{52} — 12-μμfd. tubular air condenser (see text).
C_{53} — 75-μμfd. variable (National PSE-75).
C_{54} — 125-μμfd.-per-section variable (National TMS-125-D).
C_{55} — 0.001-μfd. 2500-volt mica.
C_{57} — 470-μμfd. mica.
C_{58} — 470-μμfd. 2500-volt mica.
C_{59} — 300-μμfd. variable (National STH-300).
C_{60}, C_{61} — 0.1 μfd., 250 volts (Sprague Hypass).
R_1, R_6 — 22,000 ohms, 1 watt.
R_2 — 0.15 megohm, ½ watt.
R_3, R_{12} — 0.1 megohm, ½ watt.
R_4, R_{16} — 0.47 megohm, ½ watt.
R_5, R_9 — 1000 ohms, ½ watt.
R_7, R_{19} — 470 ohms, 1 watt.
R_8 — 1 megohm, ½ watt.
R_{10}, R_{15} — 0.22 megohm, ½ watt.
R_{11} — 12,000 ohms, 1 watt.
R_{13} — 0.5-megohm potentiometer.
R_{14} — 10,000 ohms, ½ watt.
R_{17} — 390 ohms, ½ watt.
R_{18}, R_{20} — 47,000 ohms, ½ watt.
R_{21} — 47 ohms, ½ watt.
R_{22} — 22,000 ohms, ½ watt.
R_{23} — 330 ohms, 1 watt.
R_{24} — 5000 ohms, 10 watts.
R_{25} — 1500 ohms, 1 watt.
R_{26} — 2000 ohms, 10 watts.
R_{27} — 15,000 ohms, 10 watts.
L_1 through L_{18} — See coil table.
I_1, I_2 — Panel lamp.
J_1, J_2 — Coaxial-cable connector.
J_3 — 5-prong male plug.
MA_1 — 0–25 d.c. milliammeter.
MA_2 — 0–300 d.c. milliammeter.
RFC_1, RFC_2, RFC_4, RFC_5, RFC_6, RFC_7 — 2.5-mh. r.f. choke.
RFC_3 — 300-μh. r.f. choke (Millen 34300).
RFC_8, RFC_9 — 7-μh, r.f. choke (Ohmite Z-50).
S_1 — 6-pole 3-section 5-position selector switch (Centralab 2525).
S_2 — S.p.d.t. center-off toggle switch.
T_1 — 6.3-volt 6-amp. filament transformer (Thordarson T-21F11).

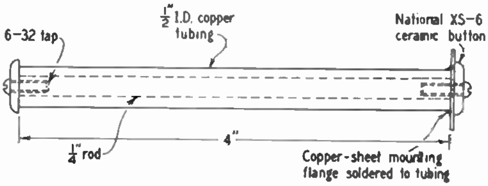


Fig. 6-44 — Sketch of the tubular condenser used in the output circuit of the bandpass all-band exciter.

the 829-B grid circuit, while the front wafer handles switching for the 7-Mc. doubler and the output links. Compensating capacitors for the doubler grid circuits are mounted between the switch sections and ground.

Rectangular cut-outs are required for the bandpass couplers. The couplers in the output of the 6AG7 and those for 28 and 27 Mc. are mounted from left to right at the rear of the chassis. The coupler for 14 Mc. is to the left of the switch, and bandpass circuits for the VFO tube and the 7-Mc. doubler are to the left and right of the aluminum partition. This aluminum shield prevents instability caused by coupling between the low-frequency circuits. L_9, the VFO coil, is mounted on a ¼-inch pillar at the front of the chassis.

The audio and the 100-kc. oscillator circuits are grouped at the upper left-hand corner and the filament transformer is bolted to the left wall. Shielded wire for all leads carrying other than r.f. will provide additional r.f. bypassing and help to reduce TVI.

Making the Band-Pass Couplers

The accompanying coil table gives the details of the various coupler windings. The dimensions listed should be followed with the greatest possible care. Especially the spacing between the two windings of the couplers is critical if the desired bandpass characteristic is to be obtained. All except the 28-Mc. coils; L_1 and L_2, make use of Millen Type 45000 1-inch-diameter coil forms. The low-frequency coupler coils, L_7 through L_{13}, are close-wound on the outside of the form in the conventional manner, except that one of the two coils in each case is made so that its position on the form can be changed slightly if necessary. This is done in the following manner. Dust the form with talcum powder. Wrap a band of Scotch tape, *adhesive side out*, around the form. Wind the required number of turns over the band of tape. The tape will hold the turns intact while the coil is removed from the form and coated with coil dope. Replace the winding on the form when the dope is dry. After final adjustment, a drop of cement will hold the winding in place on the form.

B & W self-supporting Miniductor windings are used for the 14- and 27-Mc. couplers (L_1 through L_6). These are of such diameter that when the polystyrene supporting strips are sandpapered down slightly, the coils will slide snugly inside the Millen coil forms. The forms are slotted diametrically by making a longitudinal hacksaw cut down through the center of the form to within a half inch or so of the bottom. The leads from the coils ride up and down in these slots and are cemented in place after the final adjustment of the coupling.

No form is used for the 28-Mc. coupler, L_1 and L_2.

The coupler tuning condensers are spaced out evenly on strips of ⅛-inch bakelite or polystyrene, 1¾ inches wide. As shown in the top-view photograph of Fig. 6-43, and discussed earlier, some of the couplers are grouped together on one strip. The one nearest the panel with 6 condensers is 6 inches long, the one with 4 condensers is 4 inches in length and the small piece is 2 inches long. The condensers are fastened to the strips with small machine screws through holes in the condenser tabs. Underneath, the Millen 1-inch forms are fastened to the strips with a machine screw through the hole in the bottom of the form, midway between associated tuning condensers. In the case of the 28-Mc. coupler, the coils are soldered directly to the condenser-mounting screws, in a horizontal position. Once the correct spacing has been found, the two coils are made rigid by joining them with a strip of polystyrene bridging the tops of the coils and held fast with cement.

Placement of this coupler can be seen in the bottom-view photograph of Fig. 6-45. The coils are mounted at right angles to those of the 11-meter coupler immediately to the rear of the bandswitch.

COIL TABLE FOR BANDPASS TRANSMITTER

Coil	$L_{\mu h.}$	*Wire*	*Turns*	*Diam., In.*	*Length, In.*	*Coil Spacing, In.*	*B & W Type No.*
L_1, L_3	1.18	20 tinned	8	¾	½	L_1, L_2 — ½	3011
L_2	0.99	20 tinned	7	¾	7/16	——	3011
L_4	0.81	20 tinned	6	¾	⅜	L_3, L_4 — 7/16	3011
L_5	4.1	24 tinned	15	¾	15/32	L_5, L_1 — 9/16	3012
L_6	2.26	24 tinned	10	¾	5/16	——	3012
L_7	15.8	30 enam.	21	1	7/32	L_7, L_8 — 7/16	——
L_8	9.8	26 enam.	16	1	9/32	——	——
L_9	92.0	30 s.s.c	68	1	29/32	——	——
L_{10}, L_{11}	52.5	30 enam.	42	1	7/16	L_{10}, L_{11} — 5/16	——
L_{12}	53.5	30 enam.	44	1	½	L_{12}, L_{13} — ¼	——
L_{13}	42.0	30 enam.	37	1	13/32	——	——
L_{14}	0.1	14 enam.	3	⅜	½	——	——
L_{15}	1.05	14 enam.	3	2½	⅜	——	3906
L_{16}	2.0	12 enam.	5	2½	1*	——	3905
L_{17}	6.5	14 enam.	10	2½	1¼	——	3906
L_{18}	5.4	14 enam.	9	2½	1⅛	——	3906

*** End turn adjustable — see text.**

Output Coils

The coils in the output circuit of the 829-B and their link coils are made from strip-coil material of larger size. The two coils are mounted by their leads to small stand-off insulators, with their axes at right angles. One end turn of L_{16} is broken away from the others by severing all but the bottom insulating strip. This permits the turn to be bent away from the rest of the coil for accurate adjustment of the inductance.

Tuning the Couplers

The driver stages should be adjusted in sequence, starting with the lowest-frequency band. The high-voltage supply to the 829-B should be turned off and the lead to the screens of this tube should be disconnected temporarily. With the bandswitch in the 3.5-Mc. position, set C_{34} at minimum capacitance and adjust C_{25} until the oscillator signal is heard at slightly above 4000 kc. Then the oscillator should tune over the range of about 3350 to slightly above 4000 kc.

To adjust the first coupler in the output of the oscillator, connect a high-resistance voltmeter across R_{22}, move the two coils as far apart as possible and adjust C_{41} and C_{44} for maximum meter reading. After both circuits have been peaked, do not disturb the settings of the two condensers. Slide the movable coil to give the coil spacing shown in the coil table and then check the voltmeter reading as the VFO is tuned through its range. The reading should stay constant within 10 or 15 per cent across the band. If the reading is high near both ends and low in the middle of the band, this indicates that the coupling is too tight and the coils should be moved slightly farther apart. On the other hand, if the readings show a peak in the center of the band and the excitation drops off too much at the ends of the range, the coupling is too loose and the coils should be moved slightly closer together. If the reading is much higher at one end of the band than at the other, this can usually be corrected by very slight adjustment of the tuning condensers. Do not change the condenser settings appreciably however, or it may upset the shape of the curve as a whole.

VOLTAGE AND CURRENT TABLE FOR THE LOW-LEVEL TUBES OF THE BANDPASS TRANSMITTER

Tube	*Freq., Mc.*	E_p	E_s	E_g	E_k	I_k*, Ma.*
6AG7	1.7	150	150	—12	—	8
6AG7	3.5	300	150	—35	6	20
6N7	7.	300	—	—100	13	22
6N7	14.	300	—	—70	21	20
6N7	27.	300	—	—100	14	29
6N7	28.	300	—	—65	14	30

The same procedure is followed in adjusting the coupler in the output of the 6AG7 amplifier, but this time grid current to the 829-B can be used as the indicator. The grid current should average 16 to 18 ma.

Next turn the bandswitch to the 7-Mc. position and connect the voltmeter across R_{18}. A 2.5-mh. r.f. choke should be used in series with the voltmeter lead connecting to the grid. Now adjust C_{28} until the meter readings follow the previous characteristic pattern. Then loosen the coupling between L_7 and L_8 as much as possible. Set the VFO to the middle of the designated frequency range of this stage and tune the two circuits to resonance as described previously, using the 829-B grid current as the indicator. Bear in mind that this stage and the succeeding stages need cover with essentially flat output only that frequency range for which they are labeled in Fig. 6-42. The 14-, 27- and 28-Mc. couplers are adjusted following the same procedure. The voltmeter should be connected across the grid leak only in these last three stages and no r.f. choke for the meter lead is necessary.

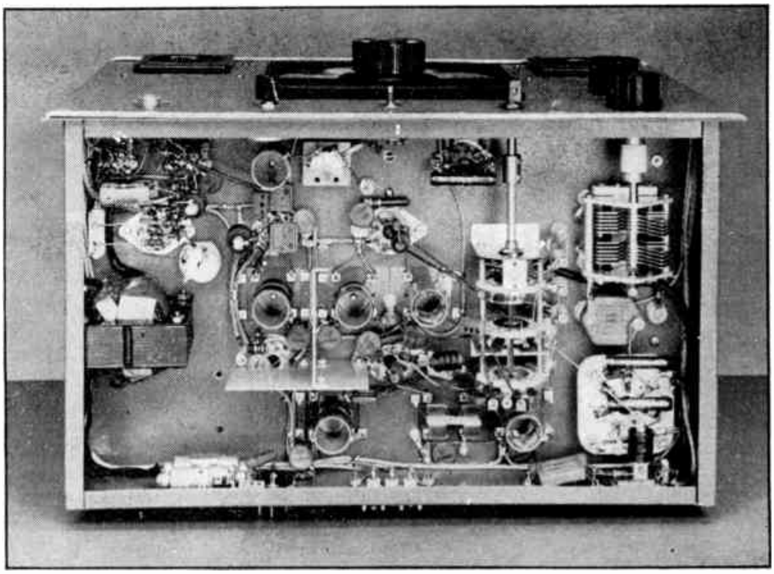


Fig. 6-45 — Bottom view of the all-band bandpass exciter. The placement of parts is discussed in the text.

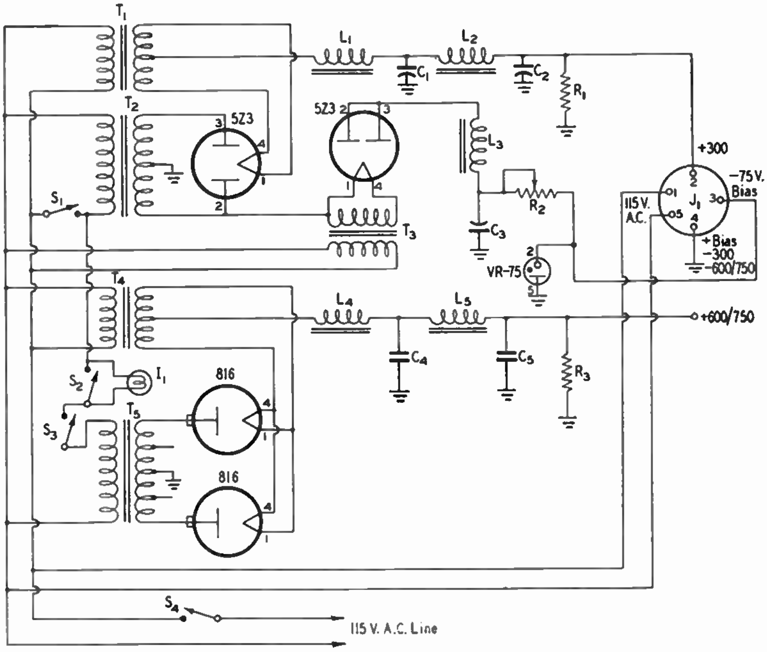


Fig. 6-46 — Circuit diagram of plate and bias supply for the all-band bandpass transmitter.

C_1, C_2, C_3 — 8-μfd. 450-volt-wkg. electrolytic.
C_4, C_5 — 4-μfd. 1000-volt oil-filled.
R_1 — 25,000 ohms, 25 watts.
R_2 — 25,000 ohms, 25 watts, with slider.
R_3 — 25,000 ohms, 50 watts.
L_1 — 5/25-h. 150-ma. swinging choke.
L_2 — 20-h. 150-ma. smoothing choke.
L_3 — 30-h. 40-ma. filter choke.
L_4 — 5/25-h. 200-ma. swinging choke.
L_5 — 10-h. 200-ma. smoothing choke.
I_1 — 150-watt 115-volt lamp.
J_1 — 5-prong socket.
S_1, S_3, S_4 — 10-amp. toggle switch.
T_1, T_3 — 5-volt 3-amp. filament transformer.
T_2 — 300-v. d.c. 150-ma. plate trans.
T_4 — 2.5-volt, 1-amp. filament transformer.
T_5 — 600/750 v.d.c. 200-ma. plate transformer.
VR — VR-75 voltage-regulator tube.

Adjusting the Output Amplifier

Since there are no coils to change in the output tank circuit, the tuning of this stage is quite simple. With the bandswitch turned to the desired band, and reduced plate voltage applied to the plate after reconnecting the screen lead, the dual tank condenser is simply tuned for the characteristic dip in plate current indicating resonance. The circuit tunes to 3.5 and 14 Mc. with the condenser near maximum capacitance. Resonance at 7, 27 and 28 Mc. will be found near minimum capacitance. However, it is important that 3.5 Mc. and 14 Mc., and also 7 Mc. and 28 Mc. do not fall at exactly the same settings, since this condition may result in excessive fourth-harmonic output when the transmitter is working at the lower of the two frequencies. The condition can be avoided by adjusting the free turn on L_{16}.

A current and voltage table shows the approximate operating conditions for the low-level tubes. Under full load, the 829-B grid current and grid voltage should average 12 ma. and 70 volts, respectively, and the screen should draw about 30 ma. at 200 volts. The amplifier may be loaded to a plate current of 200 ma.

Testing the Audio Section

The power amplifier should be turned off (do not forget to remove screen voltage) while the audio system is undergoing the first test. After a microphone has been connected to J_1 and the low-voltage supply turned on, the output signal of the transmitter should be monitored by means of a receiver. Modulation should be applied for this test and, with the receiver tuned to an n.f.m. band, the deviation control should be adjusted for a clean-sounding well-modulated signal. It must be remembered that this adjustment holds for one band only and that the deviation control requires readjustment when the transmitter is switched to another band. Less deviation is needed for the higher-frequency bands. More extensive information on aligning n.f.m. units is given in the chapter on frequency and phase modulation.

Total cathode current for the two audio tubes is approximately 1.5 ma. and about 0.5 volt is developed across the cathode resistor of each stage. Plate voltage for the speech-amplifier tube is roughly 30 volts and 25 volts should be measured at the screen-grid pins of both 6AK5s.

Testing the 100-Kc. Oscillator

Power for the 100-kc. crystal oscillator may be obtained only by turning on the transmitter supply. However, the transmitter can be disabled during the test by opening the key. S_2 must be switched to the crystal position and a receiver should be tuned to a harmonic of the crystal. A short antenna connected to the oscillator-output terminal at the rear of the chassis may be necessary if the receiver is tuned to a high frequency and if the transmitter is enclosed in the cabinet. When the circuit appears to be working normally, the oscillator may be brought to zero beat with one of the WWV frequencies by means of C_7.

Plate and screen potentials for the 6SH7 should be 150 and 50 volts, respectively. One volt should appear across the cathode resistor, R_5, and the cathode current is 1 ma.

A Shielded 150-Watt Transmitter for Four Bands

Figs. 6-47 through 6-55 show the circuit and various constructional details of a 150-watt transmitter with a shielding enclosure made of screening. In the circuit diagram of Fig. 6-48, the oscillator is a modified Pierce. It drives either a single 807W as a straight amplifier, or two of them as push-push doublers. When a single tube is used, the heater of the other 807W is turned off (S_1) and the idle tube then serves as a neutralizing condenser for the other. Type 807W tubes permit a more compact arrangement with shorter leads, but standard 807s may be substituted with very minor modifications.

To minimize v.h.f. harmonic radiation, link coupling, instead of capacitive coupling, is used between the two stages and simple harmonic filters are inserted in the power and keying leads which are shielded. R_3, R_4 and RFC_5, RFC_6 are v.h.f. parasitic suppressors.

Both stages have parallel plate feed. Since the entire unit is designed to operate from a single power supply, VR tubes are used to stabilize the plate and screen voltages of the oscillator. The 807W screen voltage also is taken from the tap for c.w. operation. If screen and plate modulation is contemplated, individual series screen resistors directly from the high-voltage terminal must be used. Each should have a rating of 50,000 ohms, 5 watts for operation at a plate voltage of 600. A small 45-volt biasing battery (90 volts for 'phone) mounted under the chassis serves to hold the amplifier input to a safe level when the oscillator is keyed. Meters with r.f. by-passes are provided in the amplifier grid and plate circuits.

VFO input can be used by means of capacitive coupling through a coaxial line and a plug (Millen 37412) that fits the crystal socket. The outer conductor of the coaxial line is grounded as close as practicable to the 6AG7 socket. When the plug is inserted, C_1 is grounded and serves as the screen by-pass condenser for the 6AG7, while the grid is connected to the "hot" side of the VFO output.

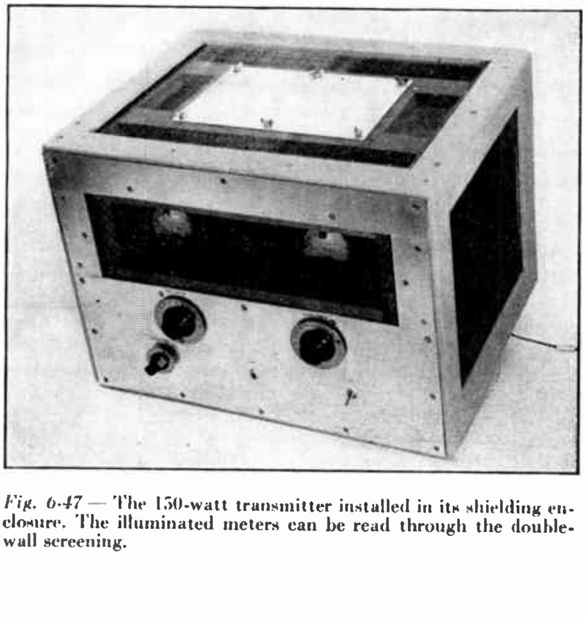


Fig. 6-47 — The 150-watt transmitter installed in its shielding enclosure. The illuminated meters can be read through the double-wall screening.

Shielding Enclosure

Screening makes a desirable type of enclosure for an amateur transmitter, since it not only provides the necessary ventilation, but also visibility. The transmitter is built in quite conventional form on a standard chassis, and the enclosure is made simply of adequate dimensions to surround it completely. The box is provided with a metal panel in front and a terminal board of the same material at the rear. The control shafts are extended the necessary distance to the panel, while the power leads are extended to the terminal board. The shielding and filtering of meters is no problem because they also are completely within the enclosure, the screening permitting reading without cut-outs.

Double-wall shielding is provided, since it is considerably more effective than single-layer screening, even though the walls are not insulated at all points.

The sketch of Fig. 6-50 illustrates the manner in which the enclosure is contructed. Each side (also top and bottom) consists basically of a square or rectangular frame of 1 by 2 pine strip stock covered with bronze screening. Copper is better if it is available. At the frame corners, the two pieces are simply butted and joined with metal angles from the dime store. To cover the edges of the frames as well as the openings, the first piece of screening is cut exactly to the *width* of the frame and about four inches longer than the length. Then one edge of the screen is tacked along the *front* face of the top strip. The screen is bent backward around the adjacent edge, stretched across the back of the frame, pulled around the opposite edge and tacked along the front face of the bottom strip. The second layer of screening is cut to a width equal to the *length* of the frame, and is applied to the *front* side of the frame in the same manner as described above, except that it is wound around the frame in the opposite direction, i.e., from side to side, instead of from top to bottom. The result is a frame that is completely covered with screening, including the edges.

Fig. 6-51 suggests a method of stretching the screening tightly across the frame. After tacking one end of the screening to the frame, the loose end of the screening is clamped between angle irons in a vise. The top strip of the frame rests against the face of the vise. When

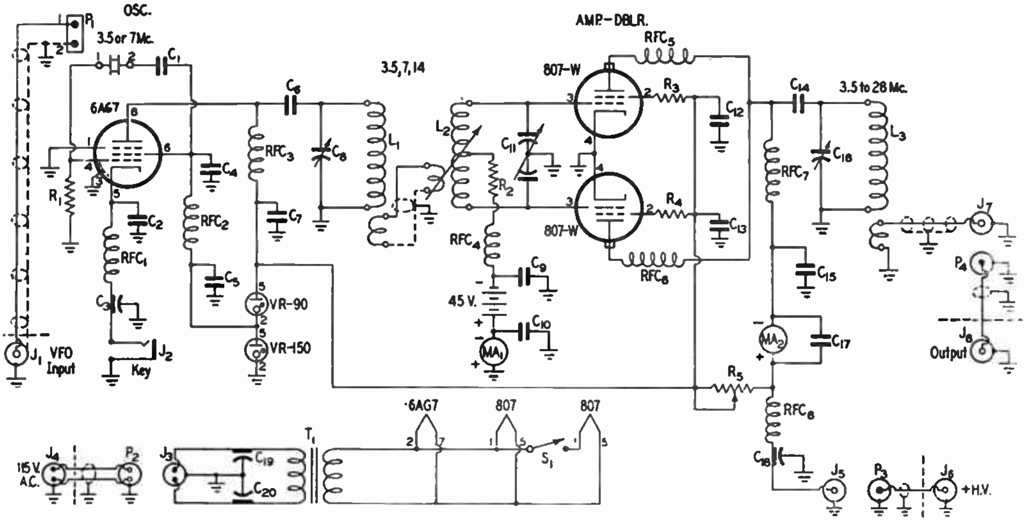


Fig. 6-48 — Circuit diagram of the shielded transmitter.

C_1, C_6 — 0.002-μfd. mica.
C_2, C_5, C_7, C_9, C_{12} C_{13} — 0.01-μfd. ceramic disk.
C_3 — 0.01-μfd. feed-through (Sprague 47P6).
C_4 — 25-$\mu\mu$fd. mica.
C_8 — 100-$\mu\mu$fd. variable (National ST-100).
C_{10}, C_{17} — 0.001-μfd. mica.
C_{11} — 100-$\mu\mu$fd.-per-section variable (National STHD-100).
C_{14}, C_{15} — 0.002-μfd. 2000-volt silicone (Plasticon ASG13 Glassmike).
C_{16} — 300-$\mu\mu$fd. variable (National TMS-300).
C_{18}, C_{19}, C_{20} — 0.005-μfd. feed-through (Sprague 46P8).
R_1 — 0.1 megohm, ½ watt.
R_2 — 1000 ohms, 1 watt.
R_3, R_4 — 100 ohms, ½ watt, noninductive.
R_5 — 15,000 ohms, 25 watts.
L_1 — 3.5 Mc. — 20 μh. — 30 turns No. 22 d.s.c., 1½ inches diam., 1½ inches long, 4-turn link (National AR17-80-E with 26 turns removed).
— 7 Mc. — 10 μh. — 18 turns No. 22 d.s.c., 1½ inches diam., 1½ inches long, 4-turn link (National AR17-40-E with 10 turns removed).
— 14 Mc. — 5 μh. — 12 turns No. 22 d.s.c., 1½ inches diam., 1 inch long, 3-turn link (National AR17-20-E).
L_2 — 3.5 Mc. — 40 μh. — 38 turns No. 22 d.s.c. close-wound, 1½ inches diam., approx. 5-turn link over center (National AR17-80-S).
— 7 Mc. — 10 μh. — 20 turns No. 22 d.s.c., 1½ inches diam., 1½ inches long, approx. 4-turn link over center (National AR17-40-S).
— 14 Mc. — 4.7 μh. — 10 turns No. 22 d.s.c., 1½ inches diam., 1 inch long, approx. 3-turn link over center (National AR17-20-S).
L_3 — 3.5 Mc. — 14 μh. — 26 turns No. 18, 1⅞ inches diam., 2½ inches long, 5-turn link (B & W JEL-40).
— 7 Mc. — 3 μh. — 10 turns No. 18, 1⅞ inches diam., 2 inches long, 3-turn link (B & W JEL-20).
— 14 Mc. — 1.5 μh. — 8 turns No. 14, 1⅞ inches diam., 2 inches long, 3-turn link (B & W JEL-10).
— 28 Mc. — 0.75 μh. — 4 turns No. 14, 1⅞ inch diam., 1 inch long, 2-turn link (B & W JEL-6).
J_1, J_7, J_8 — Jones S-101-D connector.
J_2 — Open-circuit jack.
J_3, J_4 — Amphenol 80-Pc2M connector.
J_5, J_6 — Amphenol 83-1R connector.
MA_1 — Milliammeter, 25-ma. scale.
MA_2 — Milliammeter, 300-ma. scale.
P_1 — Ribbon-line plug (Millen 37412).
P_2 — Amphenol 80-MCF1 connector.
P_3 — Amphenol 83-1SP connector.
P_4 — Jones P-101-¼-in. connector.
RFC_1, RFC_8 — 7-μh. r.f. choke (Ohmite Z-50).
RFC_2, RFC_3, RFC_4 — 2.5-mh. choke (National R50).
RFC_5, RFC_6 — 8 turns No. 18, ¼-inch diam., close-wound (National R60 — 1 μh. with turns removed).
RFC_7 — 1-mh. 300-ma. r.f. choke (National R300S).
S_1 — S.p.s.t. toggle.
T_1 — Filament transformer: 6.3 v., 2 a.

the bottom of the frame is pressed as indicated, the screening is brought under tension while it is being tacked along the face of the upper edge of the frame. The remainder of the screening is then folded over the edge of the frame and tacked along the back.

The front and rear frames are constructed as shown in Fig. 6-52A. The intermediate strip is placed to come level with the top edge of the aluminum control panel, or rear terminal board, as the case may be. Both panel and terminal board should be brought tightly against the screening by the generous use of wood screws. The construction of the top frame is shown in Fig. 6-52B. The additional crosspieces make provision for an access opening for changing coils and minor adjustments. After this frame has been covered with screening as described above, the screening across the opening can be slit and bent around the edges of the hole and tacked in place underneath. Several long machine screws are spaced around the edges of the opening so that the aluminum-sheet cover can be fastened down tightly with wing nuts. The cover should overlap the opening out to the edges of the wood framework around the hole.

The sides of the enclosure are fastened to-

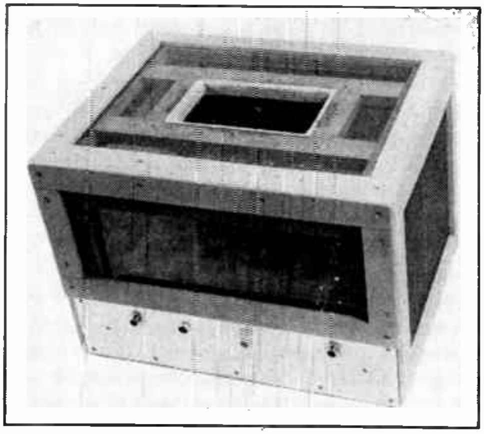


Fig. 6-49 — Rear view of the enclosure showing, from left to right, the shielded terminations for the r.f. output, a.c. line, key and VFO input. The opening in the top provides access to the plug-in coils.

gether tightly with several 1½-inch wood screws. Wood trim strips are used to cover the seams of the screening if desired. Latticing wood is suitable for this purpose.

The control shafts require holes in the two screening walls. The holes should be no larger than is necessary to pass the shafts. Ragged edges can be avoided by first flowing a small patch of solder over the screening where the hole is to be drilled, and then drilling the hole through the solder and screening.

All power and key wiring between the chassis and the terminal board at the rear should be shielded and the shield should be soldered to the screening as it passes through to the terminal board. Shielded fittings should be used as power terminals and it is advisable to use shielded wire between the terminal board and the power-supply unit.

The meters are mounted on a separate panel inside the enclosure. The panel is spaced away from the inner wall of screening by an inch or so. If there is any difficulty in reading the meters through the screening, 6.3-volt dial lamps operating from the filament transformer can be used to provide illumination for them. The lamps should be shaded toward the front to cut off glare. This can be done quite easily by coating the front of the bulbs with black paint.

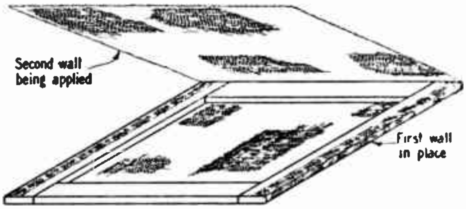


Fig. 6-50 — Two layers of screening are applied to each frame of the enclosure — one on each side of the frame.

Transmitter Construction

The amplifier tubes and their associated input and output tank circuits are constructed as a unit on a "U"-shaped bracket made from a single piece of aluminum sheet. This provides a low-inductance return, from plate circuit to cathode, independent of the chassis, as well as a measure of shielding between input and output circuits. The tank condensers are mounted directly on the bracket with their shafts at the same height. The two coil sockets are mounted above the tank condensers so that the axes of the coils are at right angles to minimize coupling. The resistors, chokes and by-pass condensers associated with the amplifier grid and screen circuits are grouped around the tube bases and connected with the shortest possible leads. Tubular-shaped C_{14} is supported (through a hole in the bracket) between the plate coil-socket terminal and the top of RFC_7 which is mounted vertically from the chassis, between the two tubes, near the plate caps. The parasitic chokes, RFC_5 and RFC_6, are suspended between the tube plate caps and the end of C_{14}. The oscillator components and the VR tubes are to the right in Fig. 6-50. The crystal, 6AG7 and the oscillator tank coil, L_1, are placed in line, with L_1 at right angles to L_2. Fig. 6-53 also shows the mounting of the two meters, and the shielded power connections.

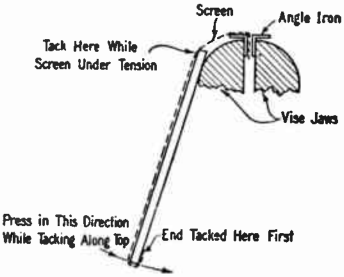
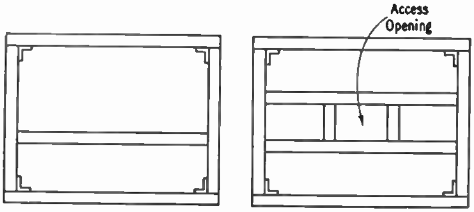


Fig. 6-51 — A suggested method for pulling the screen tight across the frame. See text.



(A) (B)

Fig. 6-52 — In the front and back frames (A) a cross-piece is added as a support for the control panel and terminal board, respectively. The top frame (B) has additional members to accommodate an access opening.

Underneath the chassis in Fig. 6-54, the oscillator tank condenser is to the left and the filament transformer and biasing battery to

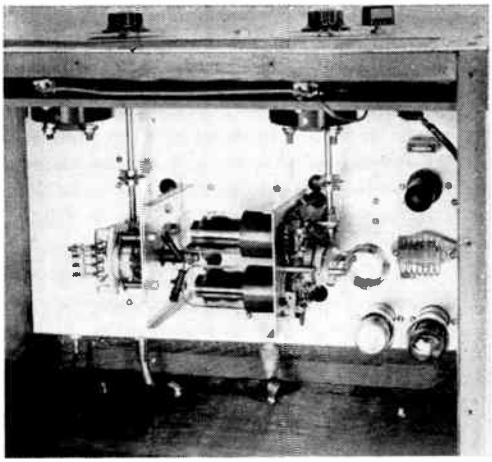


Fig. 6-53 — Top view of the 150-watt shielded transmitter. The output stage is assembled on a "U"-shaped bracket bent up from sheet aluminum. Oscillator components are to the right.

the right. R_5 is at the center. On extension leads at the top are the key jack (mounted in a National microphone-jack shield), the meter lamps, the filament switch, S_1, and the meter leads with their by-pass condensers. All power wiring is done with shielded wire with the braid shields bonded together at frequent points and tied to ground. Feed-through type by-pass condensers for harmonics are fastened directly to the a.c.-line and key terminals.

Adjustment

With the VR tubes in place, but the other tubes out of their sockets, R_5 should be adjusted until a meter connected externally in the high-voltage lead reads 40 ma. The remainder of the adjustment is quite conventional, remembering that it is possible to double frequency both in the output circuit of the oscillator and again in the output stage. Thus, output can be obtained up to the 14-Mc. band with 80-meter crystals and up to 28 Mc. with 7-Mc. crystals. C_8L_1 and $C_{11}L_2$ should always be adjusted to the same frequency.

When using VFO, it is preferable to have the VFO output one band lower than the band to which the 6AG7 output will be tuned. This avoids possible instability in the 6AG7 stage.

For maximum rated c.w. output, a 750-volt 300-ma. power supply is required (600 volts for 'phone). But a lower-voltage supply may be used for less than full output. If the supply voltage falls much below 400, however, the VR tubes will not operate, unless lower-voltage VRs are used, thus reducing the oscillator and screen voltages. Fig. 6-55 shows a suitable power supply for maximum rated operation.

The Type 807W tubes seem to work best with less than the usually-recommended grid current of 3 to 4 ma. per tube. If the grid current is run much above 2 ma. per tube, the screen current becomes excessive. At lower plate voltages, even less grid current may become desirable. At maximum plate voltage, the loaded plate current should be limited to 100 ma. per tube. At lower plate voltages, it may not be possible to load the amplifier to maximum rated plate current. In this case, the loading should be adjusted for maximum obtainable output. If antenna-current meters are not available, the amplifier should not be loaded beyond the point where there is still an easily-identifiable dip in plate current.

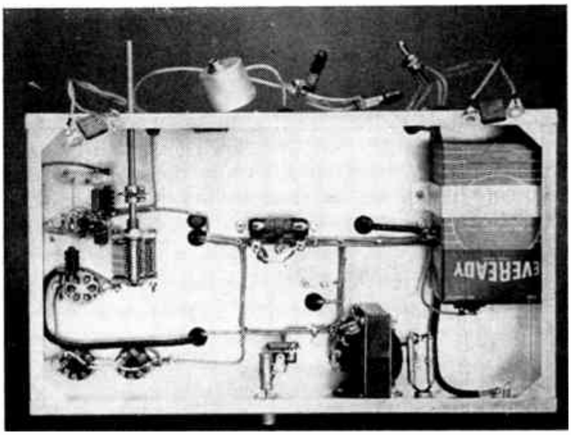


Fig. 6-54 — Bottom view of the 150-watt shielded transmitter. The chassis is of aluminum measuring 17 × 10 × 3 inches. All power wiring is done with braid-covered wire. The biasing battery to the right is held in place with a metal bracket.

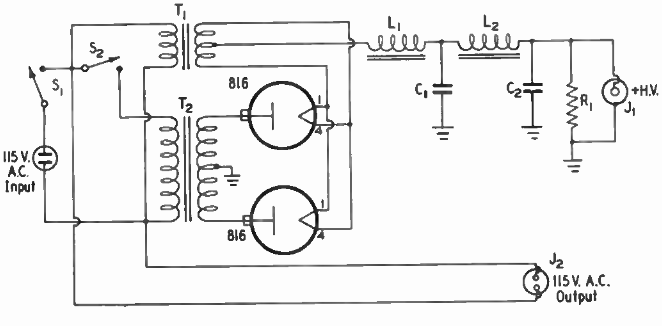


Fig. 6-55 — Circuit of a power supply for the shielded 150-watt transmitter.

C_1, C_2 — 4-μfd. 1000-volt oil-filled.
R_1 — 25,000 ohms. 50 watts.
L_1 — 5/25-h. 300-ma. swinging choke.
L_2 — 10-h. 300-ma.
J_1 — Amphenol 80-PC2F Connector.
J_2 — Amphenol 83-1R connector.
S_1, S_2 — 3-amp. toggle switch.
T_1 — Filament transformer: 2.5 volts, 4 amp.
T_2 — Plate transformer: 600/750 volts d.c., 300 ma.

A 500-Watt Link-Coupled All-Band Transmitter

In the design of the transmitter shown in Figs. 6-56 through 6-67, an attempt has been made to incorporate means by which harmonic radiation 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. 6-58, 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 "neutralizing condenser" for the other 807.

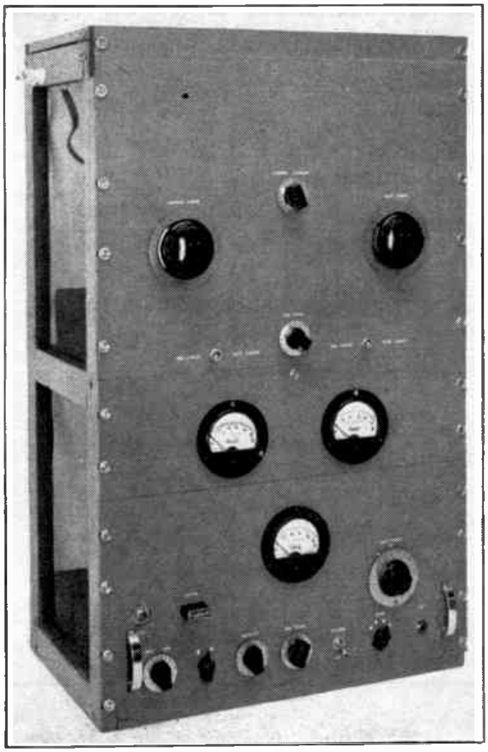


Fig. 6-56 — A complete 500-watt all-band transmitter including antenna tuner. The exciter unit at the bottom slides out for coil changing. The panel screws on this unit are dummies cemented in place. The top of the screened enclosure is hinged to permit changing coils in the final amplifier and antenna tuner.

Fig. 6-57 — 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.

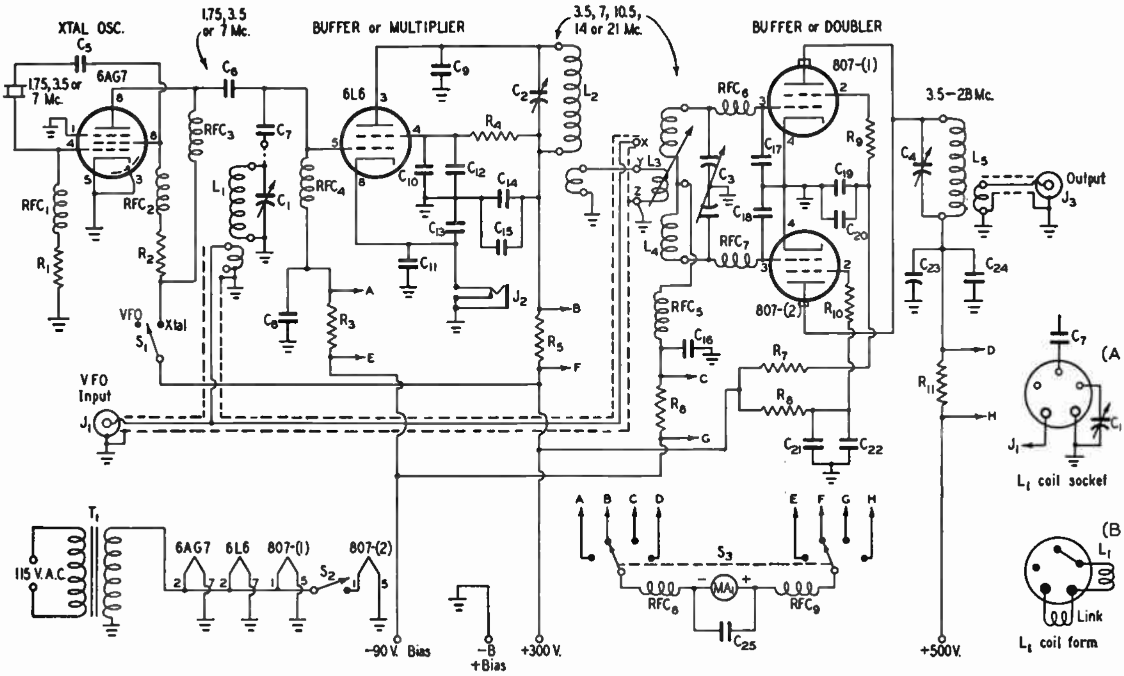


Fig. 6-58 — Circuit diagram of the exciter for the 500-watt all-band transmitter.

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 / 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

When the output frequency desired is the same as the crystal frequency, the input circuit of the 6L6 is not tuned. 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. 6-58A 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 coil base can be wired to make this connection when the coil is plugged into place.

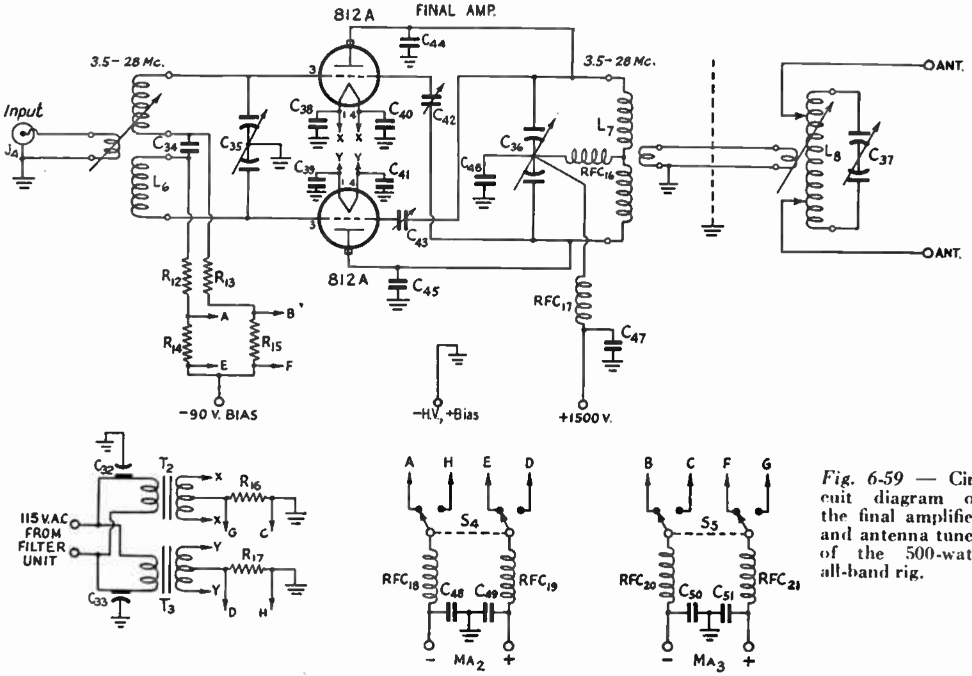


Fig. 6-59 — Circuit diagram of the final amplifier and antenna tuner of the 500-watt all-band rig.

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} — 470-μμfd. mica.
C_{42}, C_{43} — Neutralizing condenser — 4–14 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 812As).
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.

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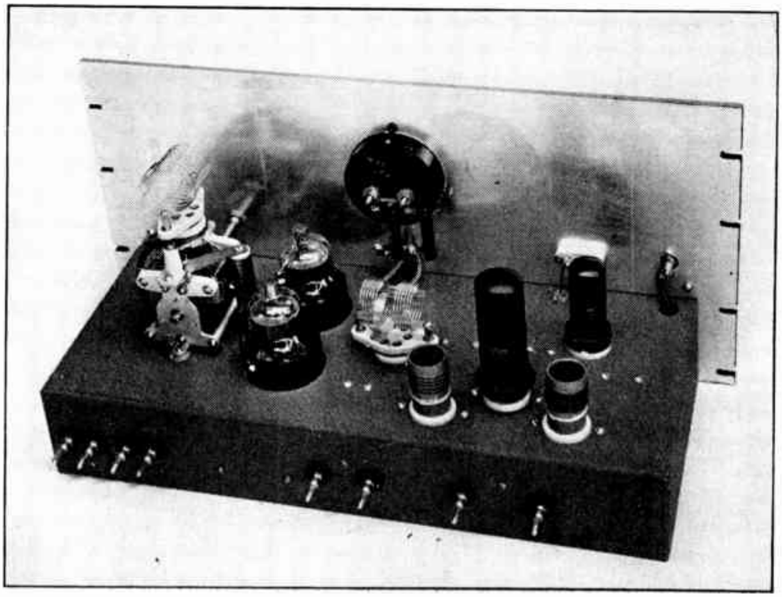

Fig. 6-60 — Rear view of the exciter of the 500-watt all-band transmitter.

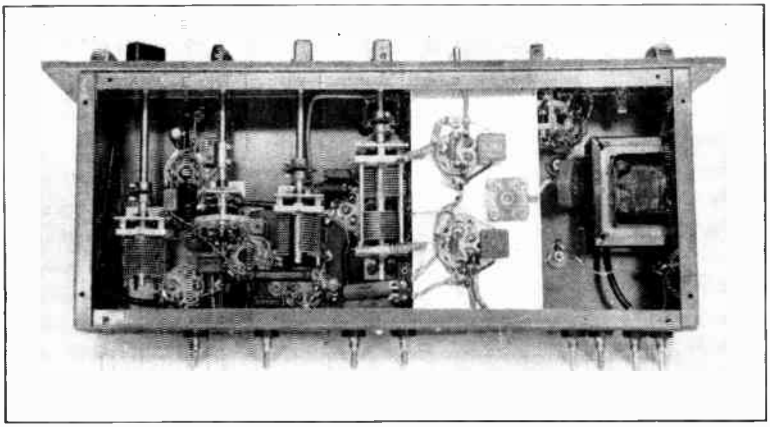

Fig. 6-61 — Bottom view of the exciter section of the 500-watt all-band transmitter. The filament transformer is to the right.

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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 in the chapter on measuring equipment. 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. 6-59. 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 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. 6-65.

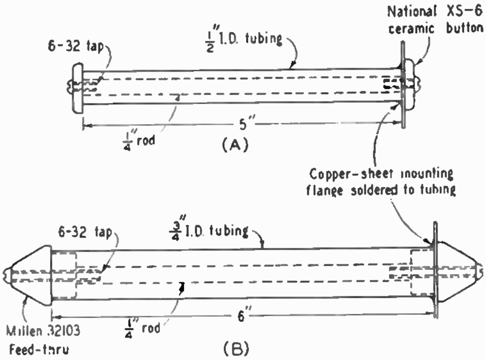

Fig. 6-63 — Sketches showing the construction of the air tubular condensers, (A) for the exciter and (B) for the final amplifier. The smaller condensers shown in the photographs of the final were replaced after preliminary tests.

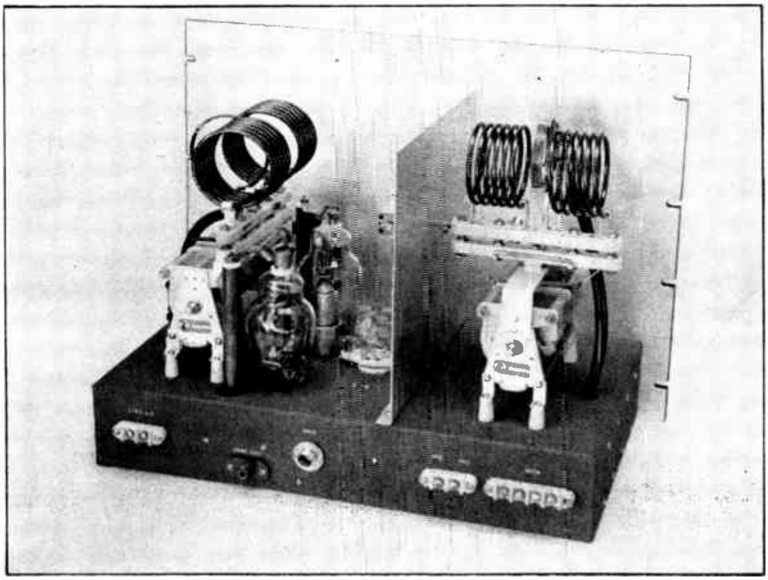

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Fig. 6-62 — Rear view of the amplifier section of the 500-watt all-band transmitter. The antenna tuner is to the right. The gear box driving the variable link of the antenna tuner is fastened to the left-hand side of the shielding partition.

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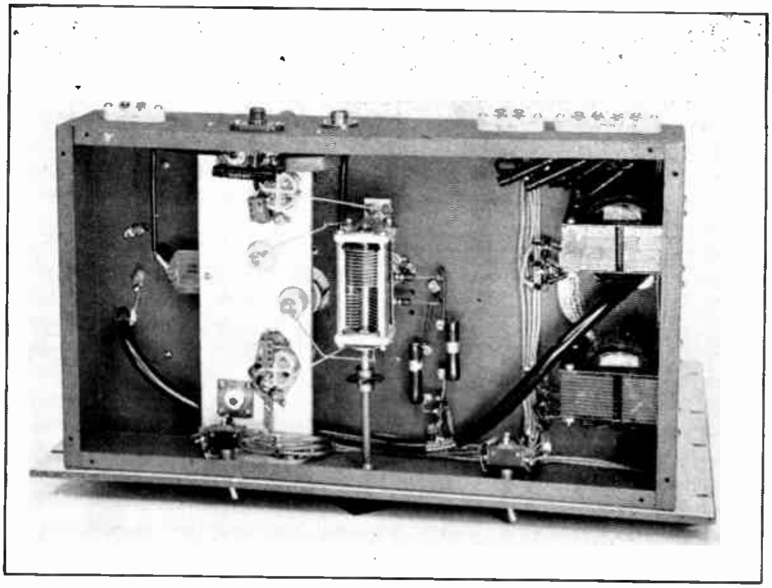

Fig. 6-64 — Bottom view of the amplifier section of the 500-watt all-band transmitter. The large coaxial lead runs from the amplifier link to the antenna-tuner link. The two filament transformers are to the right.

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. 6-60. 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.

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. 6-63A. In the push-push stage, the grid tank condenser, C_3, is im-

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	31	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	40	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-17-10-S
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″	1	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

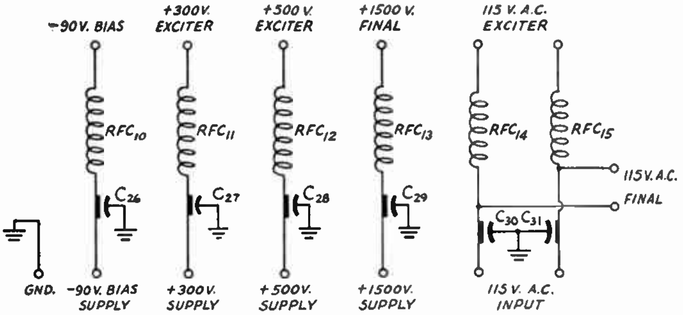

Fig. 6-65 — Wiring diagram of the harmonic-filter unit for the 500-watt all-band transmitter.

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).

mediately to the left of the tube sockets in Fig. 6-61 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 crystal socket is mounted on the panel where it is readily accessible. Connections to it are made by way of feed-through points.

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 by-pass grounding point.

Figs. 6-62 and 6-64 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. 6-63B.

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.

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 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 is shown in Fig. 6-67.

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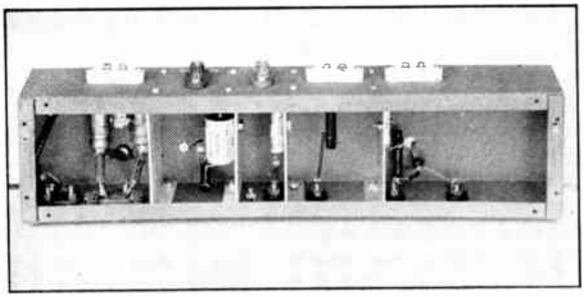

Fig. 6-66 — Bottom view of the line-filter unit. The chassis is divided off into shielded compartments by aluminum partitions. Left to right, the filters are for the a.c. line, 1500-volt d.c. line, 500-volt d.c. line, 300-volt d.c. line, and bias.

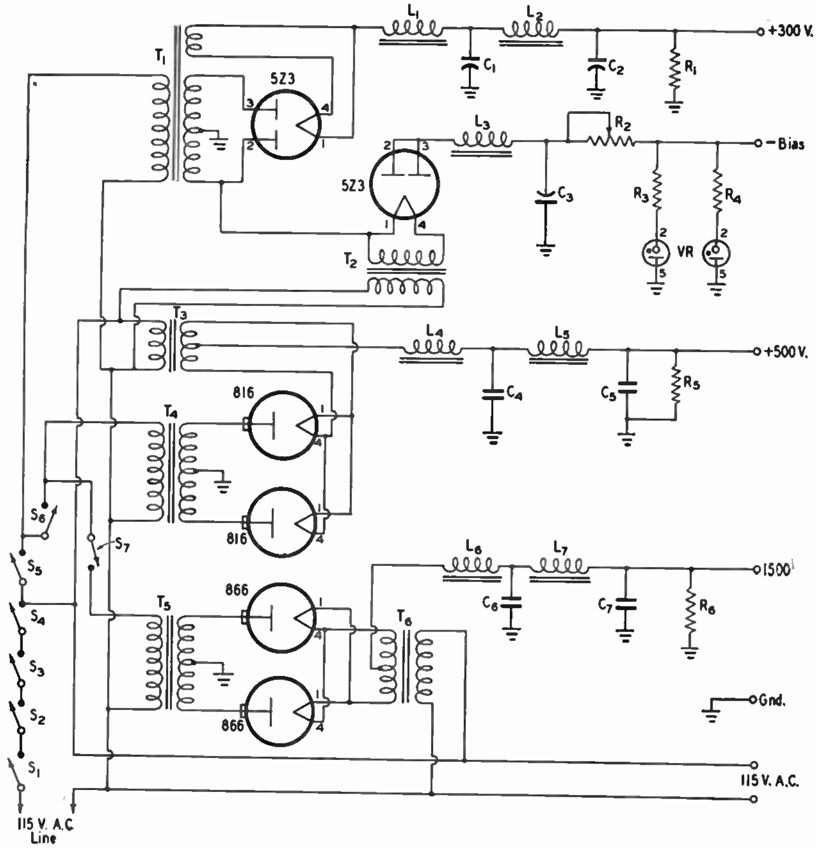

Fig. 6-67 — Circuit diagram of a power supply for the 500-watt all-band transmitter.

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. 6-59) must be changed to suit (see "R.F. Power-Amplifier-Tube Operating Factors," this chapter).

A Push-Pull 813 Transmitter

Shown in Figs. 6-68 through 6-75 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. 6-69. 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.

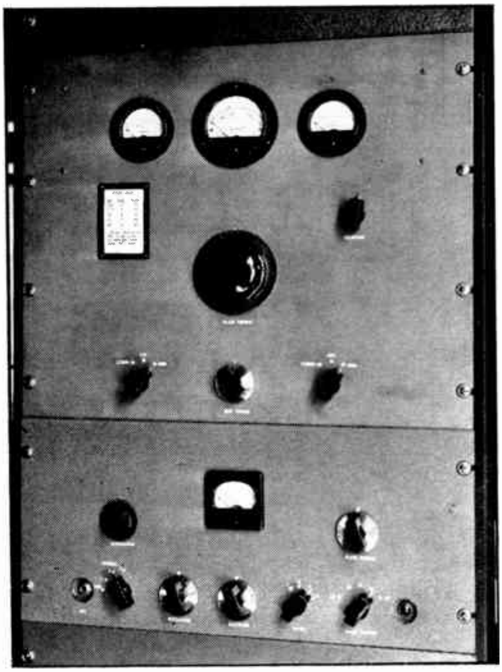

Fig. 6-68 — Front view of the push-pull 813 amplifier and its exciter mounted in a standard rack. From left to right, the controls along the lower edge of the exciter panel are for the crystal-VFO switch, the oscillator tuning condenser, multiplier tuning condenser, meter switch and bandswitch. The controls 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. 6-72. The amplifier is link coupled to the exciter. A multiband tuner (National MBS-30) 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.

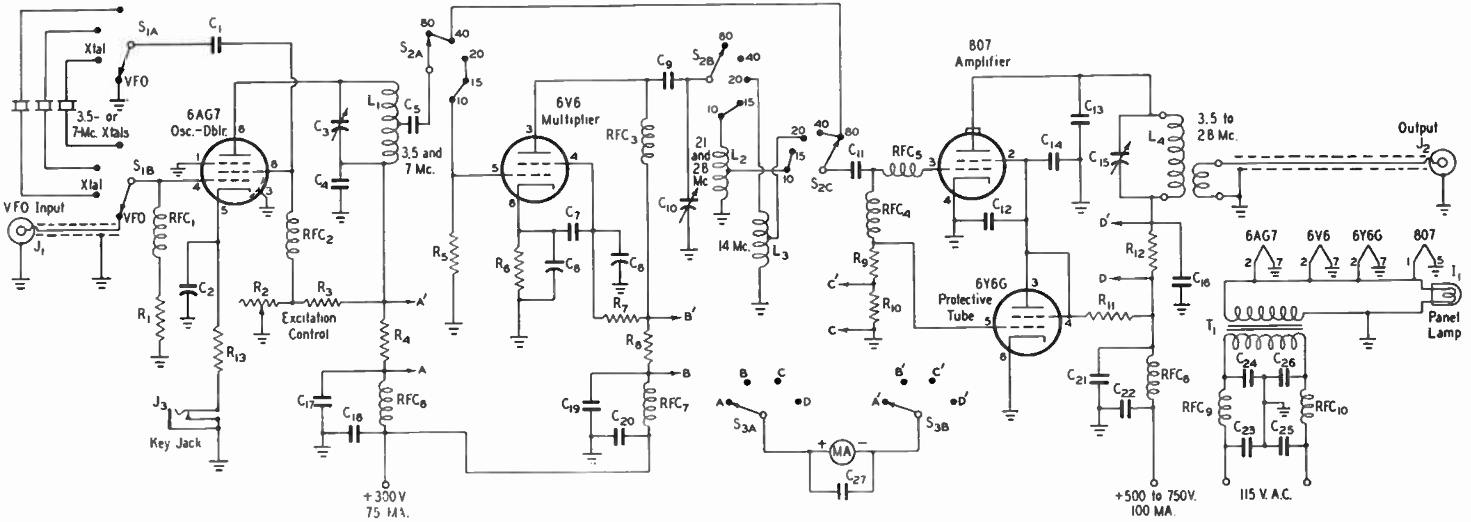

Fig. 6-69 — Circuit diagram of the exciter of the push-pull 813 transmitter.

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 43042, 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).

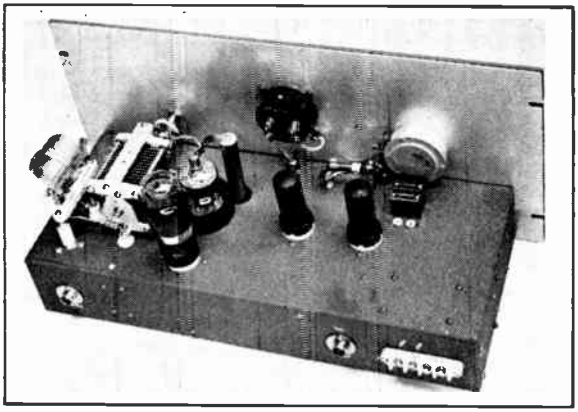

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Fig. 6-70 — Rear view of the exciter chassis for the push-pull 813 transmitter. The tubular condenser is alongside the 807 in back of the 6Y6G. The 6V6 multiplier is in the center of the chassis, with the 6AG7 oscillator, the crystal sockets and the 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. In fringe areas it may be necessary to shield the meter face.

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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. 6-63A, 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. 6-71, 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. 6-75 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

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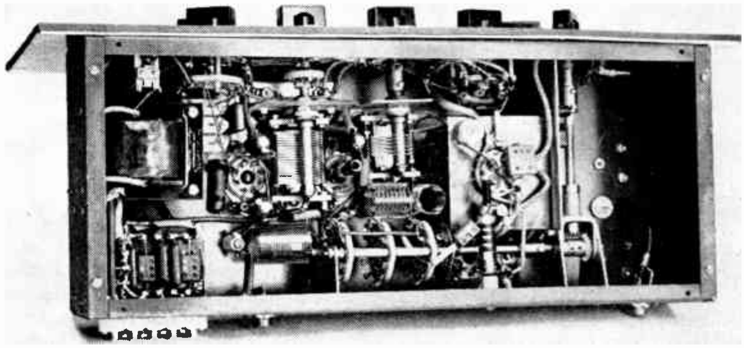

Fig. 6-71 — Bottom view of the exciter for the push-pull 813 transmitter. The harmonic filters are mounted on terminal boards placed in the lower left-hand corner, adjacent to the input terminals and just below the filament transformer. The mounting brackets for the bandswitch and the right-angle drive are supported by the rear of the chassis, while the bracket that supports the 807 socket assembly extends back from the front.

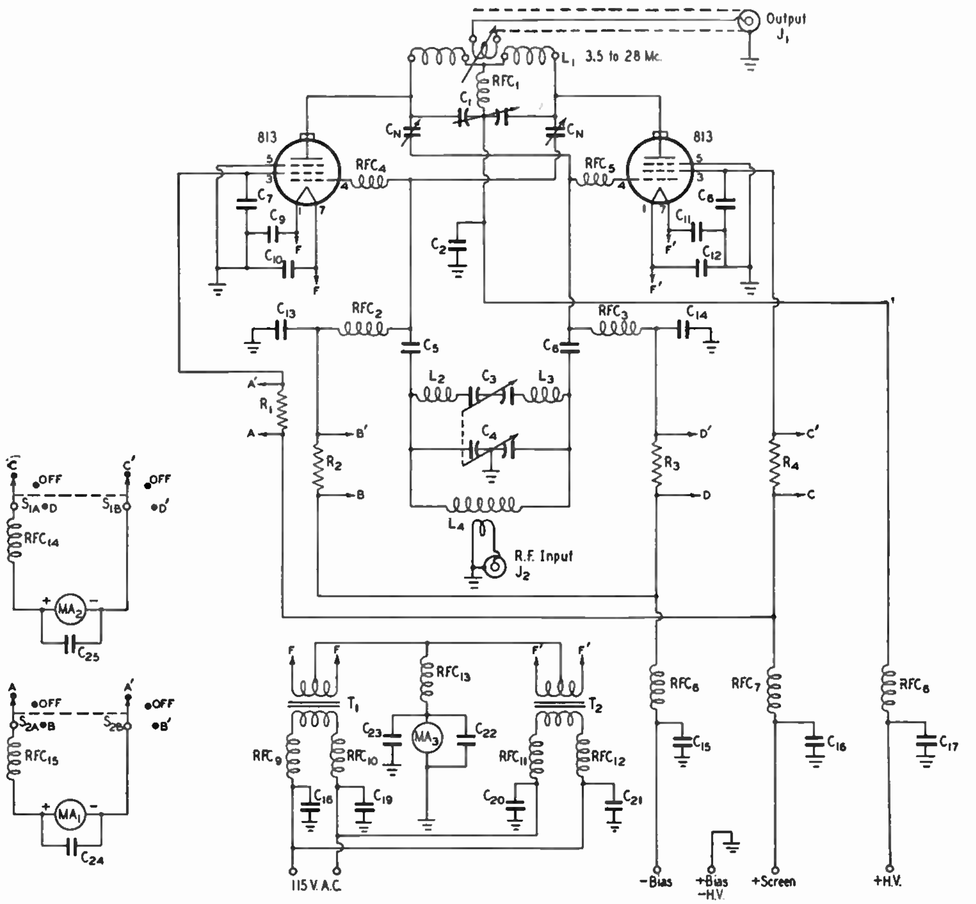

Fig. 6-72 — Schematic diagram of the push-pull 813 amplifier.

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

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Fig. 6-73 — 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.

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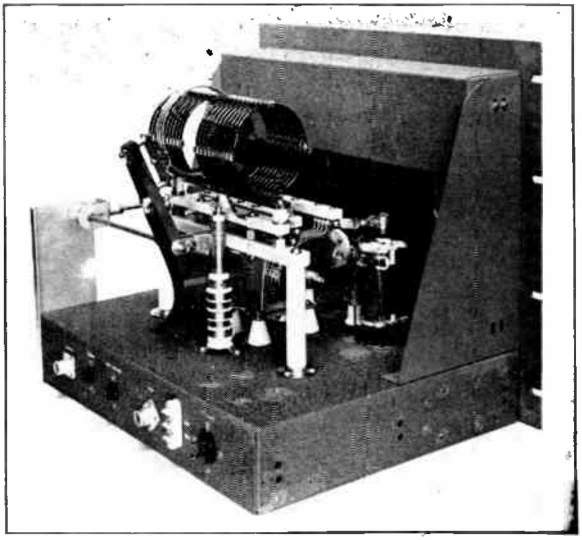

control, to the minimum that will give rated grid current to the final amplifier with optimum coupling to the 807 when the final is loaded fully. It should be possible to do this without exceeding a grid current of 3.5 ma. to the 807 and with the plate current between 60 and 90 ma.

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

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Fig. 6-74 — 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.

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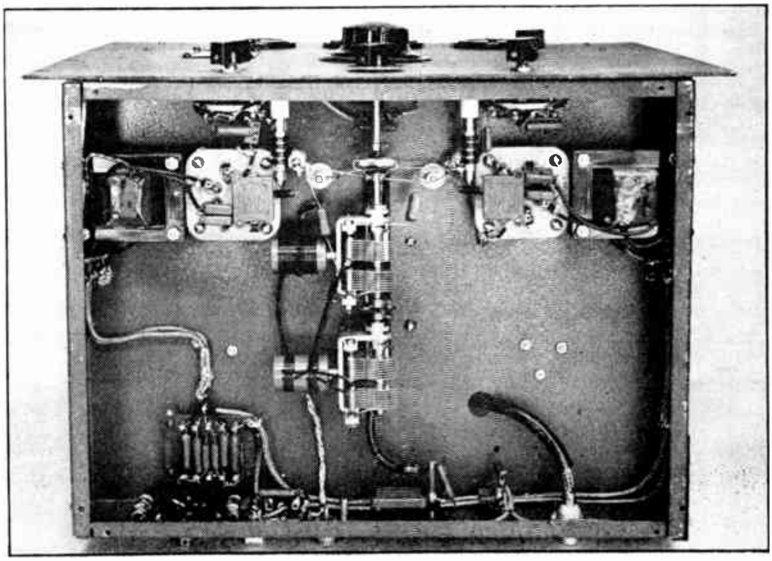

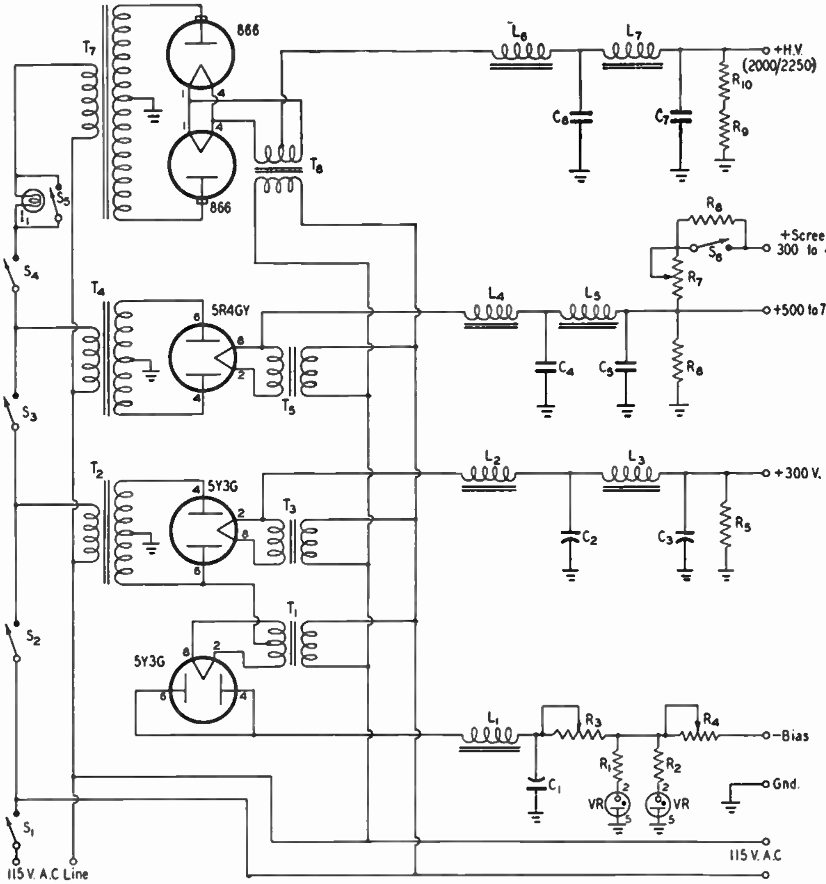

Fig. 6-75 — Circuit diagram of a power-supply system for the push-pull 813 transmitter.

C_1, C_2, C_3 — 8-μfd. 450-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: 450 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 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. 6-76 through 6-80. 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-

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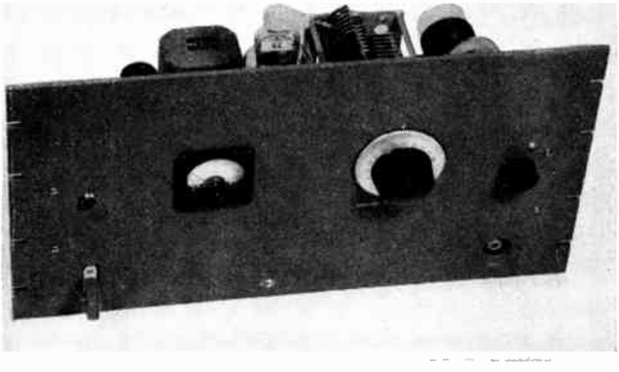

Fig. 6-76 — Front view of a 175-watt transmitter for the 160-meter band. Only one tuning control is needed, plus a small knob used to adjust the setting of the swinging link on the output coil.

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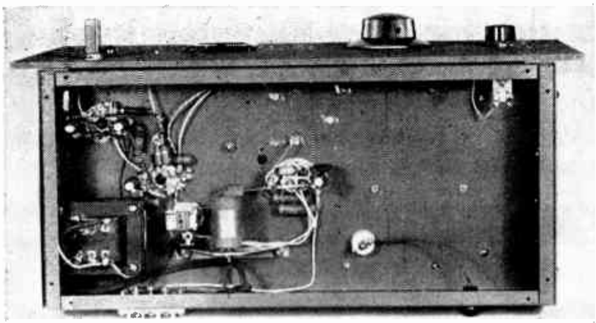

Fig. 6-77 — Bottom view of the 160-meter transmitter. The oscillator tube socket and its related parts are in the upper left corner. The 6L6 and the final-amplifier tube are mounted in a line through the center of the chassis, with the plate coil for the 6L6 supported on a bracket between the two stages. The parasitic-suppressing choke is mounted between the grid terminal of the amplifier socket and a ceramic stand-off insulator.

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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. 6-79 shows the diagram of a suitable power supply in case a separate supply is necessary or desirable. Control circuits are included.

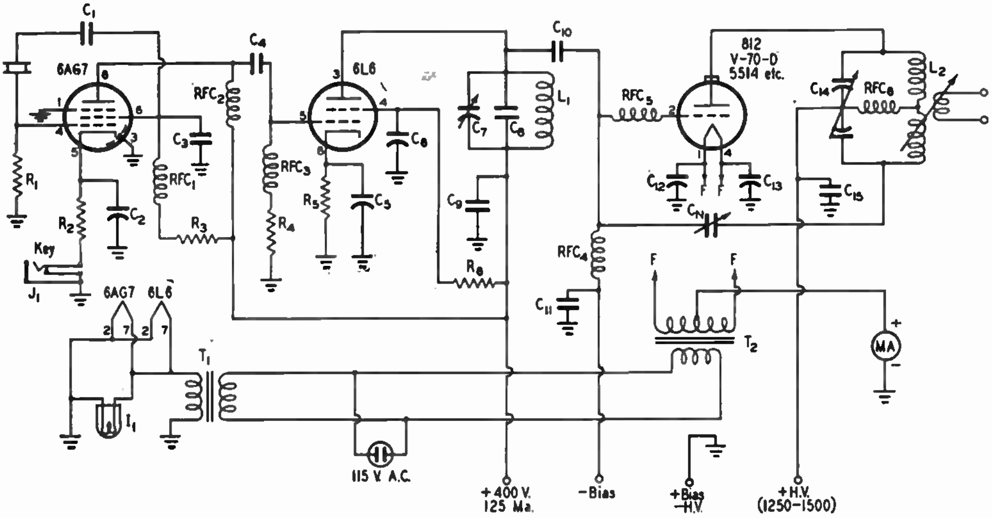

Fig. 6-78 — Schematic diagram of a single-control 175-watt transmitter for the 160-meter band.

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).

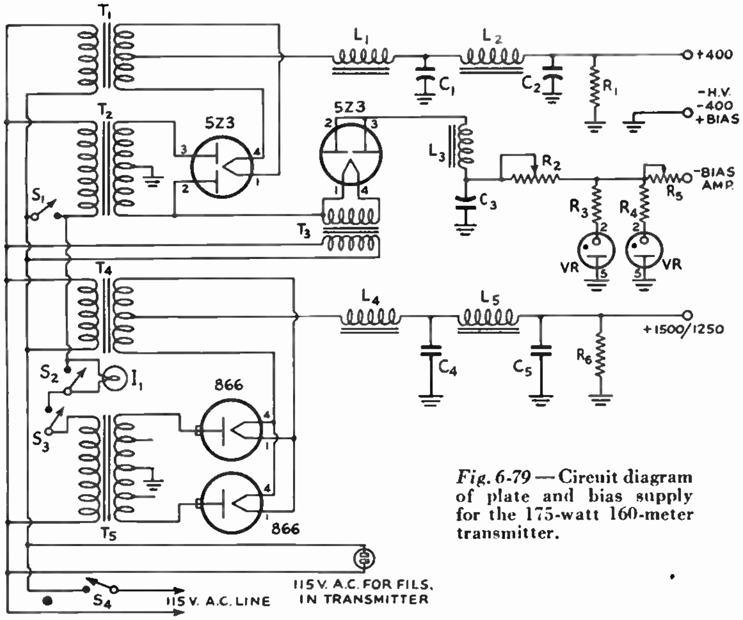

Fig. 6-79 — Circuit diagram of plate and bias supply for the 175-watt 160-meter transmitter.

C₁, C₂, C₃ — 8-μfd. 600-volt-wkg. elec.
C₄, C₅ — 4-μfd. 2000-volt oil-filled.
R₁ — 25,000 ohms, 25 watts.
R₂ — 30,000 ohms, 10 watts, with slider.
R₃, R₄ — 47 ohms, 1 watt.
R₅ — See text.
R₆ — 25,000 ohms, 150 watts.
L₁ — 5/25-hy. 150-ma. swinging choke.
L₂ — 20-hy. 150-ma. smoothing choke.
L₃ — 30-hy. 75-ma. filter choke.
L₄ — 5/25-hy. 200-ma. swinging choke.
L₅ — 10-hy. 200-ma. smoothing choke.
I₁ — 150-watt 115-volt lamp.
S₁, S₃, S₄ — 10-amp. toggle switch.
S₂ — 5-amp. toggle switch.
T₁, T₃ — 5-volt 3-amp. filament transformer.
T₂ — 400-v. d.c. 225-ma. plate transformer.
T₄ — 2.5-volt 10-amp. filament transformer, 10,000-volt insulation.
T₅ — 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.

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Fig. 6-80 — 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 Simple VFO

The details of a simple VFO with output at 1.75, 3.5 or 7 Mc. are shown in Figs. 6-81 through 6-85. In the circuit, shown in Fig. 6-84, a Type 5763 miniature pentode in a series-tuned Colpitts oscillator circuit drives a similar tube as an amplifier or doubler. The output circuit of the oscillator stage is broadbanded through the use of self-resonant slug-tuned coils at L_2, and frequency may be doubled in this circuit, as well as in the output circuit, to obtain 7-Mc. output. For 3.5-Mc. output, frequency may be doubled in either stage. The nominal output is approximately 2 watts — sufficient for driving the usual crystal-oscillator stage of the transmitter.

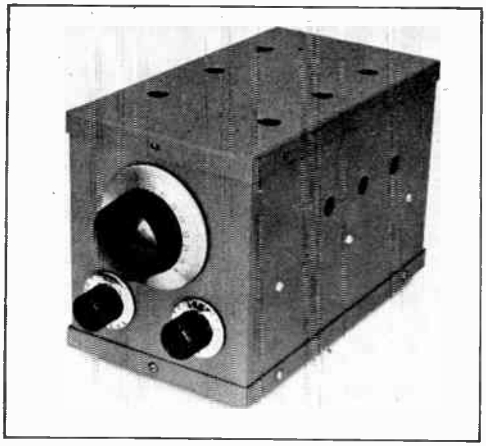

Fig. 6-81 — A simple VFO delivering output at 1.75, 3.5 or 7 Mc.

To simplify the bandspread problem, the oscillator tuning range is restricted. At 3.5 Mc. a range of approximately 250 kc. is covered. For c.w. operation in this band, the band-set condenser, C_2, is set so that the tuning condenser, C_1, covers approximately 3500 to 3750 kc. For operation in the 'phone portion of the band, C_2 is reset to shift the range to approximately 3750 to 4000 kc. Corresponding ranges are provided at the harmonics, and the oscillator can be tuned low enough (by C_2) to cover the 11-meter band with appropriate doublers.

Construction

The unit is built in a 5 × 6 × 9-inch steel box with cap-type covers. The components are assembled on an aluminum-sheet base supported by sections of aluminum angle stock that hold the base half way between the two covers. On top, the tuning condenser, C_1, is fastened directly to the base along the center line. The shaft is fitted with a National Type AM vernier dial. The two tubes and L_2 are in line to the right in Fig. 6-82, with the output tank coil, L_3, to the left of the amplifier tube. The L_2 coils are wound on Millen Type 74001 shielded slug-tuned forms.

Underneath, in Fig. 6-83, the band-set condenser, C_2, is mounted against the front of the box. A short lead through a feed-through point or clearance hole connects the stator of C_2 to the stator of C_1 above. L_1 is wound on a Millen 1-inch coil form and is placed immediately to the rear of C_2. The output tank condenser, C_{14}, is mounted on a bracket with its rear stator termi-

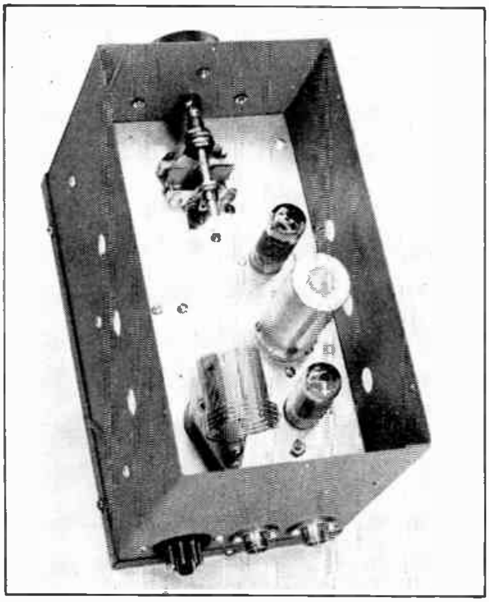

Fig. 6-82 — The top of the simple VFO showing the oscillator tuning condenser, the tubes and plug-in coils.

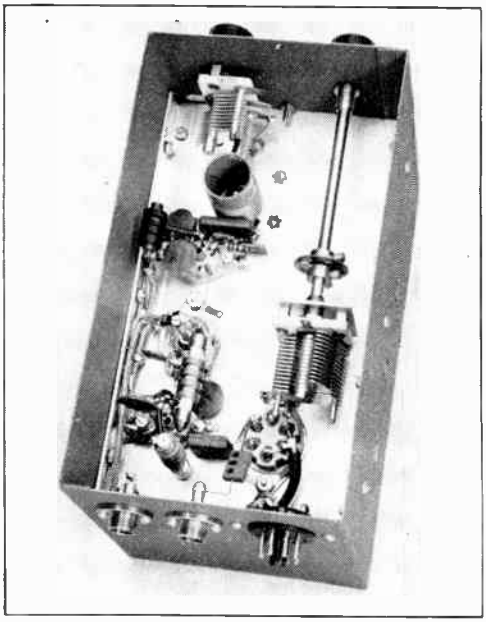

Fig. 6-83 — Bottom view of the simple VFO showing the arrangement of parts underneath.

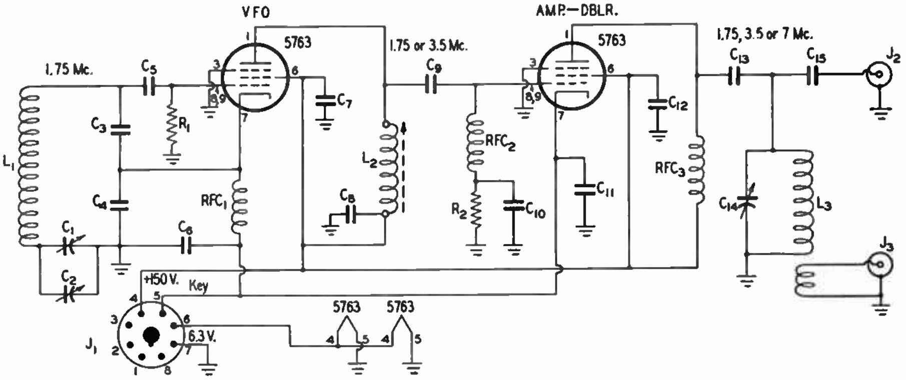

Fig. 6-84 — Circuit diagram of the simple VFO.

C_1 — Approx. 15-μμfd. variable (Millen 19025 with all but 1 rotor and 2 stators removed).
C_2 — 100-μμfd. variable (Millen 22100).
C_3, C_4 — 0.001-μfd. silvered mica.
C_5, C_9, C_{15} — 100-μμfd. mica.
C_6, C_7, C_8, C_{11}, C_{12} — 0.01-μfd. disk ceramic.
C_{10}, C_{13} — 0.001-μfd. disk ceramic.
C_{14} — 140-μμfd. variable (Millen 22140).
R_1, R_2 — 47,000 ohms, ½ watt.
L_1 — 62 turns No. 30 d.s.c., 1 inch diam., close-wound.
L_2 — 1.75 Mc. — 210 turns No. 36 d.s.c., ½ inch diam., close-wound (Millen 74001 form).(300 μh.)
— 3.5 Mc. — 126 turns No. 30 d.s.c., ½ inch diam., close-wound (Millen 74001 form). (75 μh.)
L_3 — 1.75 Mc. — 55 μh. — 45 turns No. 22 d.c.c., 1½ inches diam., close-wound (National AR17-80E).
— 3.5 Mc. — 16 μh. — 20 turns No. 22 d.c.c., 1½ inches diam., close-wound (National AR17-40E).
— 7 Mc. — 5 μh. — 12 turns No. 22 d.c.c., 1½ inches diam., ¾ inch long (National AR17-20E).
J_1 — Chassis-mounting octal plug.
J_2, J_3 — Female coaxial connector (Jones S101-D).
RFC_1 — 2.5-mh. r.f. choke (National R-50).
RFC_2, RFC_3 — 2.5-mh. r.f. choke (standard type).

nal close to the coil socket. It is placed so that its insulated shaft-extension control will balance up with the control for C_2 in front.

The various r.f. chokes and fixed condensers are grouped closely around the sockets with which they are associated in the circuit. All power wiring is done with shielded wire and coaxial output terminals are provided at the rear for either capacitive or link coupling. Key and power connections are made through the octal plug. Several ventilating holes are cut in the longer sides of the box and also in the top cover.

Adjustment

The unit requires a regulated 150-volt supply. The supply diagrammed in Fig. 6-85 is suitable. First adjust R_1, Fig. 6-85, to the maximum resistance that will permit the VR150 to stay ignited when the key is closed. Then, listening on a calibrated receiver, close the key, set C_1 at maximum capacitance and adjust C_2 until the oscillator signal is heard at 3500 kc. Tuning C_1 should then cover the band up to about 3750 kc. Mark the setting of C_2, set C_1 at maximum again and adjust C_2 until the signal is heard at 3750 kc. Then C_1 should cover the range from 3750 to approximately 4000 kc. Repeat the process, setting C_2 for about 3350 kc. to obtain the proper range for 11 meters.

To adjust the remainder of the circuit, turn the slug of L_2 in full. Touch a small neon bulb to the capacitive output terminal and adjust C_{14} for maximum indication. Check the output frequency with a wavemeter, since indications may be obtained at any multiple of 1.75 Mc. When the VFO is connected to a following stage, C_{14} and L_2 should be adjusted for maximum grid current. For capacitive output coupling, connection is made at J_2, while J_3 is provided for link coupling. With capacitive coupling, the output tank circuit should resonate with coaxial-cable lengths up to five or six feet. The frequency should be rechecked, since the setting of C_{14} will be influenced somewhat by the length of the coaxial cable with capacitive coupling. C_{14} may require an occasional touch-up in tuning the VFO across the band. A milliammeter connected in series with the key should read approximately 40 ma.; about half of this is taken by the oscillator screen and plate circuits.

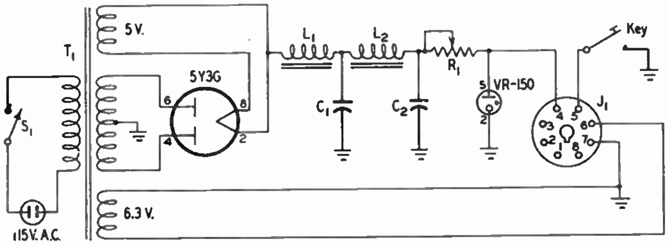

Fig. 6-85 — Circuit diagram of a power supply for the simple VFO.

C_1, C_2 — 16-μfd. 450-volt electrolytic.
R_1 — 5000 ohms, 25 watts, adjustable.
L_1, L_2 — 10-h. 50-ma. filter choke.
J_1 — Octal socket.
S_1 — 3-amp. toggle switch.
T_1 — Power transformer: 325-0-325 volts r.m.s., 40 ma.; 6.3 volts, 2 amp.; 5 volts, 2 amp.

A Silenced VFO for Break-In C.W.

Unfortunately, there is no known practical way in which an oscillator, particularly one of the VFO type, can be keyed without a compromise in respect to chirps or clicks. Steps taken to eliminate one will aggravate the other. In the VFO unit shown in Figs. 6-86 through 6-90, the oscillator is not keyed, but allowed to run continuously while a subsequent amplifier is keyed. The signal from the oscillator is suppressed by proper shielding and circuit design, so that it causes no interference to reception on any frequency, including the operating frequency, even with the receiver r.f. gain control at maximum. Any desired shaping of the keyed signal can be applied to the amplifier without introducing chirps.

A block diagram of the system is shown in Fig. 6-87 and the circuit diagram appears in Fig. 6-88. A very low-power high-C Hartley oscillator (15 to 20 volts at the plate), using a 6SK7 and operating in the region of 875 kc. drives a second 6SK7 as a strictly Class A isolating amplifier at the same frequency. The Class A stage, in turn, drives a doubler to 1750 kc. This doubler stage is keyed by the blocked-grid method. Thus, until the key is closed, most of the signal is confined to 875 kc. Further suppression of harmonics from the oscillator is obtained by omitting the cathode by-pass condenser in the buffer stage, thereby introducing a slight amount of degeneration.

The output circuits of both the oscillator and buffer are broadbanded and require only initial adjustment. The tuning controls of the oscillator and the output doubler are ganged. A suitable power supply is included on the chassis.

Construction

The photographs of Figs. 6-86 and 6-89 show one method of construction. The layout is not critical, and may be changed to meet individual needs provided that certain considerations are kept in mind. It is desirable to have as much isolation between stages as possible to eliminate stray coupling of the oscillator harmonic to the output circuit. For this reason all heater and d.c. supply leads are made with shielded wire, with the shield braid grounded at several points. The bottom of the 12 × 10 × 3-inch chassis is divided into three compartments by simple aluminum shield partitions. One, which runs the full depth of the chassis along one side, contains all the power-supply components. Another, in the front part of the chassis, contains the oscillator circuit. The third compartment contains the Class A amplifier and the output stage. The coils in these stages (L_3 and L_4) should be kept as far apart as possible, and mounted at right angles to one another. If a more compact layout is used, it is suggested that a shield be installed between them.

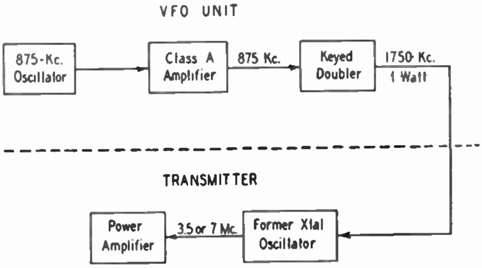

Fig. 6-87 — Block diagram of the system. The circuits so closely resemble those of the usual VFO unit that modification of an existing unit to use the new system should be simple. The underlying principles that make the new system better than the old are discussed in the text.

The wiring is arranged so that there is only one lead passing from one r.f. compartment to

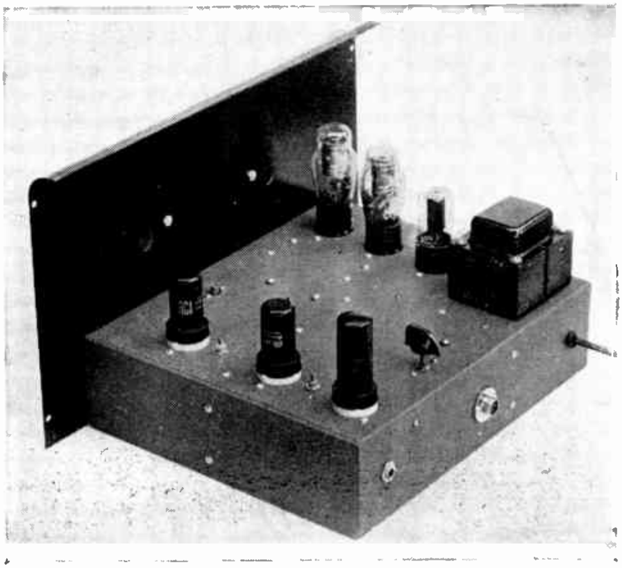

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Fig. 6-86 — Rear view of the break-in VFO unit. The oscillator tube and its padder condenser are near the panel, with the Class A amplifier and the 6AG7 output stage in line behind it. The slug-adjusting screws of the plate coils are visible between the tubes. The output padding condenser is mounted so that its shaft comes through the chassis near the 6AG7. The key jack and the coaxial connector used for a shielded output terminal are mounted on the rear of the chassis. All power-supply components are grouped along one side of the chassis.

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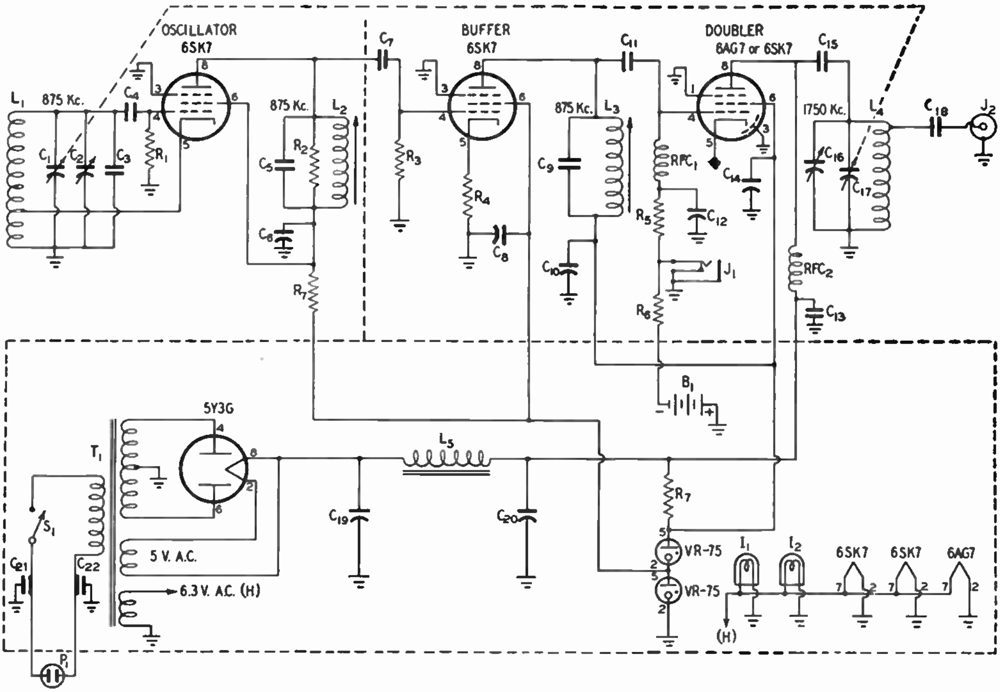

Fig. 6-88 — Schematic diagram of the VF.O

C_1 — 200-μμfd. max. variable (Millen 19200).
C_2, C_{16} — 140-μμfd. trimmer (Millen 22140).
C_3 — 680-μμfd. silver mica.
C_4, C_7, C_{11}, C_{15} — 100-μμfd. mica.
C_5, C_9 — 68-μμfd. mica.
C_6, C_8, C_{10}, C_{14} — 0.01-μfd. 600-volt paper.
C_{12} — 0.02 to 0.1-μfd. paper.
C_{13} — 0.01-μfd. disk ceramic.
C_{18} — 220-μμfd. mica.
C_{17} — 50-μμfd. variable (Hammarlund MC-50-S).
C_{19}, C_{20} — 8-μfd. 450-volt electrolytic.
C_{21}, C_{22} — 0.1-μfd. 250-volt a.c. (Sprague Hypass).
R_1, R_7 — 47,000 ohms, ½ watt.
R_2, R_3 — 10,000 ohms, ½ watt.
R_4 — 330 ohms, ½ watt.
R_5 — 0.1 megohm, ½ watt.
R_6 — 0.22 megohm, 1 watt.
R_7 — 5000 ohms, 20 watts.
L_1 — 47 turns No. 24 d.s.c. close-wound on 1-in. diam. form. Tap at 14 turns from ground.
L_2, L_3 — Slug-tuned coil (CTC LS-3, 1 Mc.). (350 μh)
L_4 — 53 turns No. 26 d.s.c. close-wound on 1-in. diam.
L_5 — 10-hy. 75-ma. filter choke.
B_1 — 67.5 to 90-volt battery.
I_1, I_2 — 6.3-volt dial lamps (part of dial assembly).
J_1 — Closed-circuit 'phone jack.
J_2 — Coaxial connector, female.
P_1 — Male a.c. plug.
RFC_1, RFC_2 — 2.5-mh. r.f. choke (National R-100-S).
S_1 — S.p.s.t. toggle switch.
T_1 — 700 volts c.t., 70 ma.; 5 v., 3 amp.; 6.3 v., 2.5 amp. (Stancor P-6011).

the other. This requires that the d.c. and heater leads to each stage be brought out separately to the power-supply compartment. The coupling lead between the plate of the oscillator and the grid of the Class A stage is the only lead passing through the partition between the two stages. In connection with the wiring in the Class A stage, the components in the grid circuit should be spaced from the plate circuit to reduce the possibility of regeneration.

In the rear-view photograph of Fig. 6-86, the r.f. tubes are lined up at the left with the oscillator tube near the panel and the 6AG7 output tube to the rear. The adjusting screws for L_2 and L_3 are between the tubes. The knob at the rear is for C_{16}. Power-supply components are to the right.

Underneath, in Fig. 6-89, the oscillator components are at the lower left, with the band-set condenser C_2, L_1 and the oscillator tube to the left of the tuning condenser, C_1. L_2 is to the rear of the oscillator-tube socket. In the compartment to the rear, L_4 is mounted horizontally from the shielding partition and the amplifier-tube sockets are to the left.

Adjustment and Operation

Some adjustment of the amount of fixed capacitance used in the oscillator circuit may be required to permit tuning the range 875 kc. to 1000 kc. If only c.w. operation is wanted, greater bandspread of the low-frequency end of the bands may be obtained by removing rotor plates from C_1, and making up the difference in fixed capacitance with C_2. A wide frequency range is possible with the constants shown, including coverage of the 11-meter band, by readjustment of C_2.

The most important adjustment is to make sure that the Class A stage is operating true Class A. To do this, resonate the plate circuit of the oscillator in the center of the desired tuning range. Then do the same for the plate circuit of

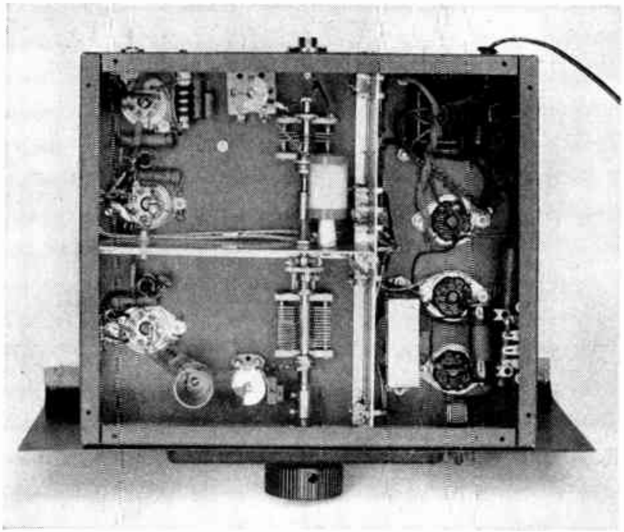

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Fig. 6-89 — **Bottom view of the silenced VFO unit. In this view, the three compartments are shown. The oscillator occupies the lower left-hand compartment, with the Class A amplifier and the 6AG7 output stage in the compartment above it. The power supply is in the compartment at the right.**

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the Class A stage. If no wavemeter capable of tuning the required range is available, a receiver tuned to the broadcast band can be used. Connect a low-range voltmeter, through an r.f. choke, across the cathode resistor, R_4, of the Class A amplifier. About 3 volts bias should be indicated. Now pull the oscillator tube out of its socket. The voltage read across R_4 will remain the same if Class A conditions are being met. If they are not, reducing oscillator output by increasing the size of R_7, or decreasing the size of either R_2 or R_3, should correct the trouble.

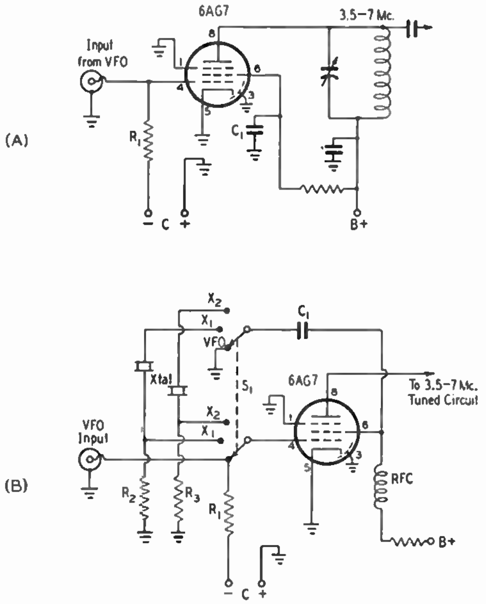

Fig. 6-90 — **Two suggested methods of coupling the VFO unit to the transmitter. In both cases the 6AG7 is used as either a doubler or quadrupler from the output of the VFO. In A, a former crystal-oscillator stage has been revised to operate with fixed bias. In B, a switching system providing for either VFO or crystal control is shown.**

C_1 — 0.001-μfd. (or larger) mica.
R_1 — 10,000 ohms, ½ watt.
R_2, R_3 — 47,000 ohms, ½ watt.
S_1 — Double-pole 3-or-more-position ceramic.

Now connect the output of the VFO to the input of the transmitter through a shielded lead. Coaxial cable, with the outer shield grounded, can be used here to good advantage, because it is low-loss and well shielded. Any length up to 8 or 10 feet can be used. The cable should be terminated at the transmitter end close to the grid of the first stage, so that radiation will not take place from any exposed wiring in the transmitter. Press the key, and adjust trimmer C_{16} for maximum grid current in the first stage of the transmitter. If too much drive is available, as well may be the case if a 6AG7 is used as the first stage in the transmitter, use a 6SJ7 output stage in the VFO unit. This will reduce the output considerably. The substitution can be made without wiring changes, but the tuning of both the Class A stage and the output stage may have to be readjusted slightly. If an 807 is used as the first stage of the transmitter, a 6AG7 will probably be needed as the output tube in the VFO, but for smaller tubes, the 6SJ7 should be adequate.

To eliminate the last trace of signal from the oscillator, it is usually necessary to apply a certain amount of fixed bias to the stage into which the VFO unit as shown is to work. Fig. 6-90 shows the method of applying bias to the usual 6AG7 crystal-oscillator circuit. Only the minimum value of bias needed to suppress the oscillator signal should be used; the balance can be made up by grid leak. Some value between 3 and 22.5 volts should be sufficient. If the battery used in the keying circuit has appropriate taps, the same battery may be used for both purposes. Small-size batteries may be used, since the drain is negligible.

A Bandswitching Gang-Tuned VFO-Exciter

Figs. 6-91 through 6-94 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. 6-93, 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. 6-92, 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

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Fig. 6-91 — The gang-tuned bandswitching exciter is built in a standard amplifier-foundation enclosure. Additional holes have been punched in the top to aid ventilation.

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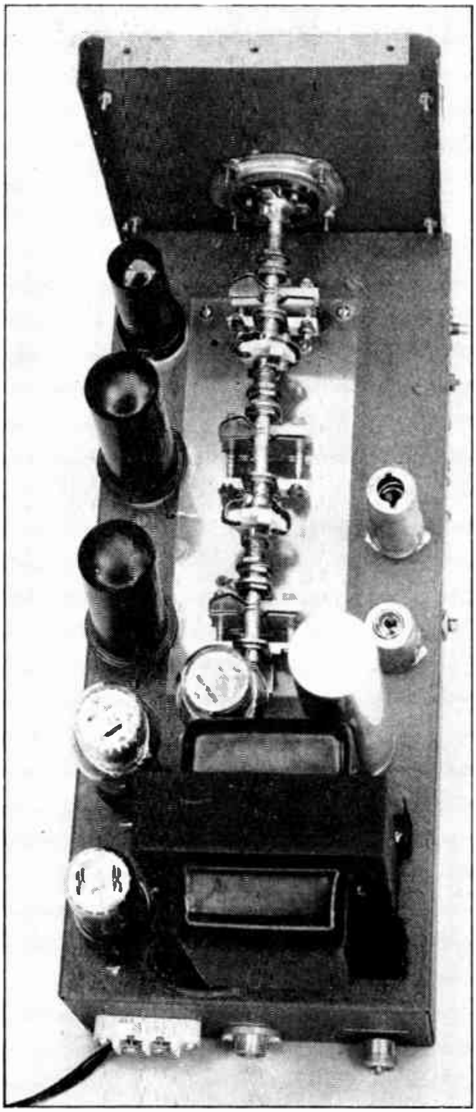

Fig. 6-92 — Rear view of the gang-tuned bandswitching exciter.

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. 6-94, 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. 6-94 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.

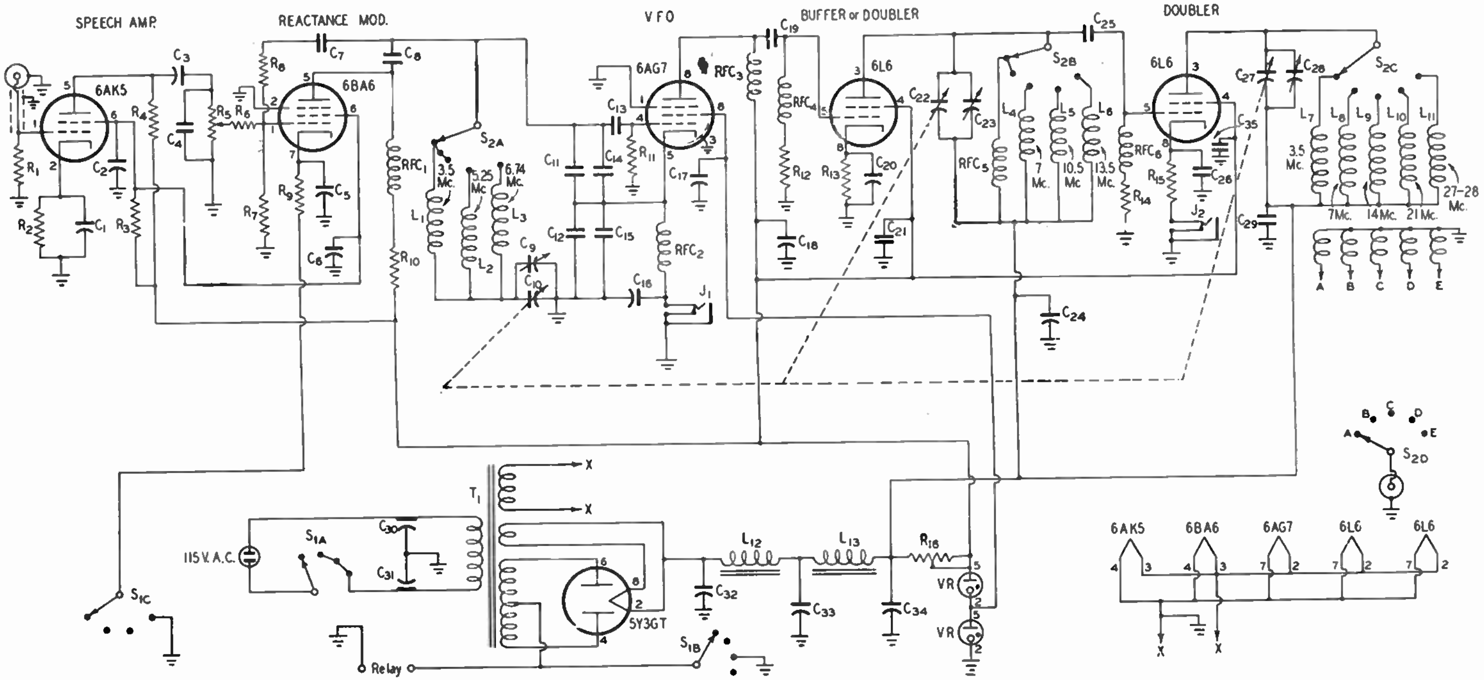

Fig. 6-93— Circuit diagram of the gang-tuned bandswitching exciter.

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-uh. 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.

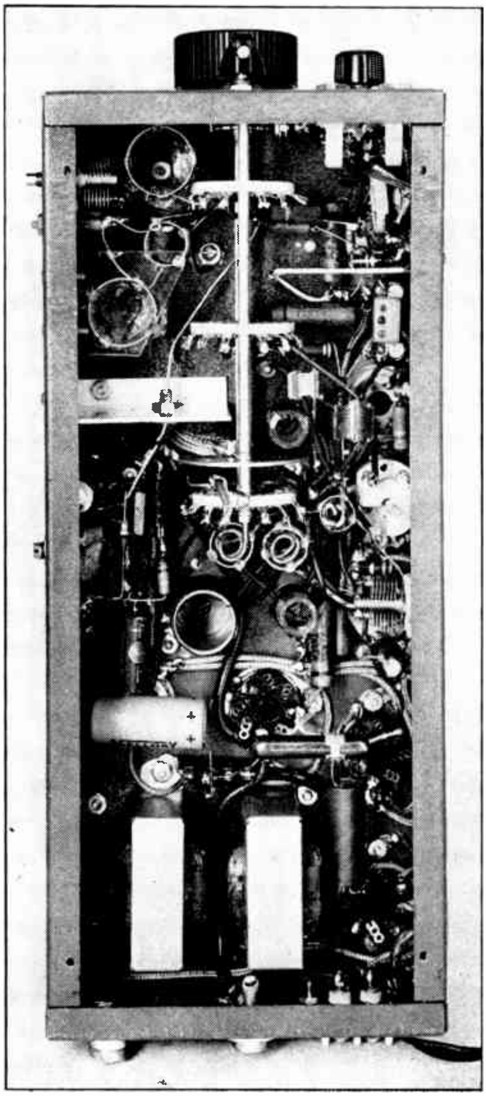

Fig. 6-94 — Bottom view of the gang-tuned bandswitching exciter. The chassis is 3 inches deep.

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 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. 6-95 through 6-100. 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. 6-97, 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.

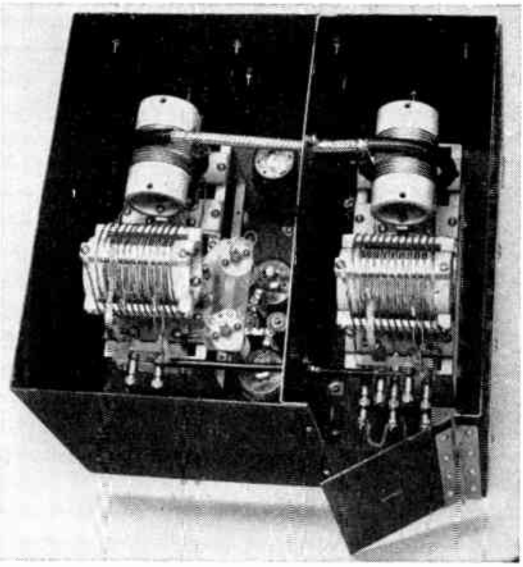

Fig. 6-96 — Rear-top view of the multiband 807 amplifier, showing the hinged door in the rear for altering antenna connections. The amplifier is to the left, the antenna tuner to the right. The third tube is the 6Y6G. The outer conductor of both link lines is grounded on the shielding partition.

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

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Fig. 6-95 — 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.

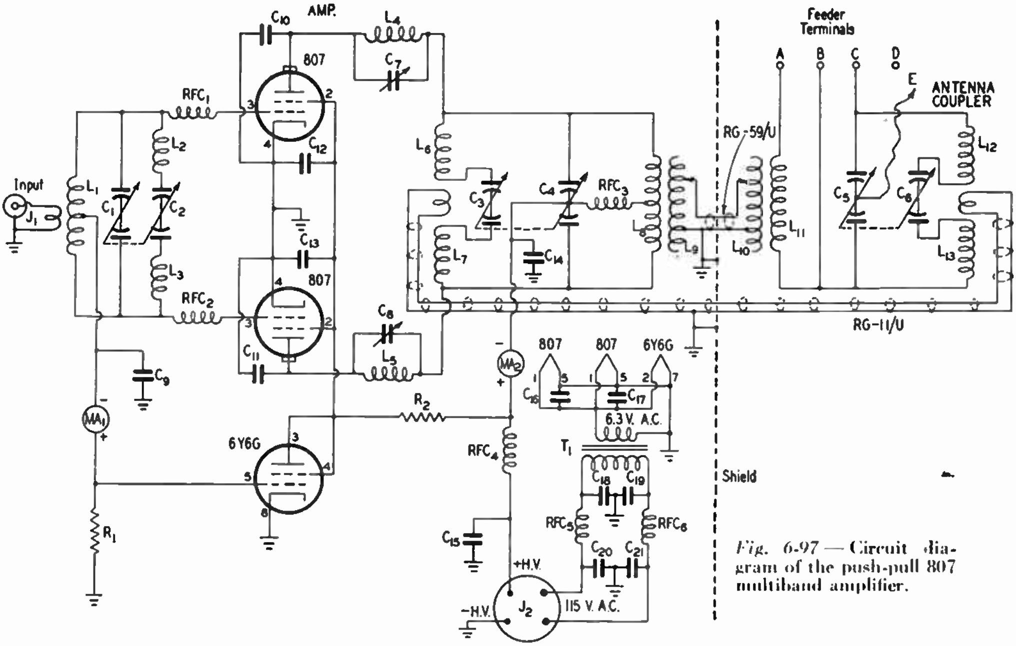

Fig. 6-97 — Circuit diagram of the push-pull 807 multiband amplifier.

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, I watt.
R_2 — 25,000 ohms, 20 watts.
L_1 — 30 turns No. 22 enam., center-tapped, $1\frac{1}{4}$ inches long, 1-inch diam.
L_2, L_3 — 7 turns No. 22 enam., $\frac{5}{16}$ inch long, 1-inch diam., with $\frac{3}{8}$-inch space between sections.
(NOTE: Above coils are part of MB-40 tuner.)
L_4, L_5 — 4 turns No. 16 tinned, 1 inch long, $\frac{5}{16}$-inch diam.
L_6, L_7, L_{12}, L_{13} — 5 turns No. 12, $\frac{5}{8}$ inch long, $1\frac{3}{4}$-inch diam., with $\frac{3}{8}$-inch space between sections.
L_8, L_{11} — 18 turns No. 12, 2 inches long, $1\frac{3}{4}$-inch diam.: L_8 is center-tapped.
L_9, L_{10} — 12 turns No. 12, $2\frac{1}{2}$ inches long, $2\frac{1}{2}$-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. 6-98. 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. 6-97. 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

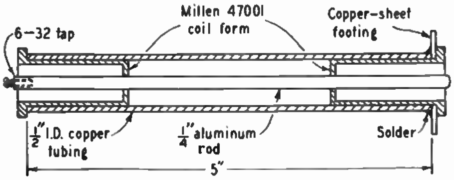

Fig. 6-98 — Sketch showing the construction of the tubular air condensers used in the p.p. 807 amplifier. The condenser is mounted with screws through the footing.

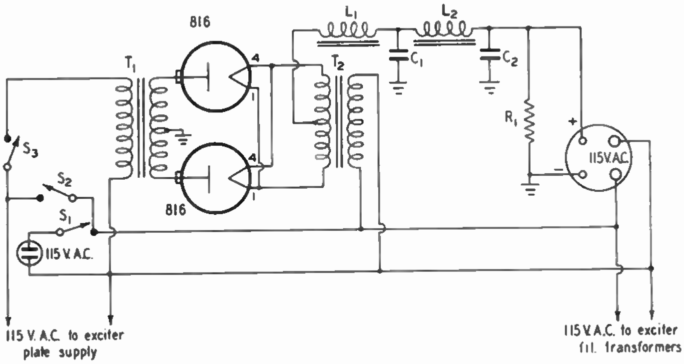

Fig. 6-99 — Circuit of power supply for the push-pull 807 amplifier.

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, 4 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.

Adjustment

The diagram of a power supply for this amplifier is shown in Fig. 6-99. 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. 6-97, 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.

The antenna tuner is adjusted following conventional procedure (see chapter on transmission lines). 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 necessity for any change in feeder connections to the antenna coupler.

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

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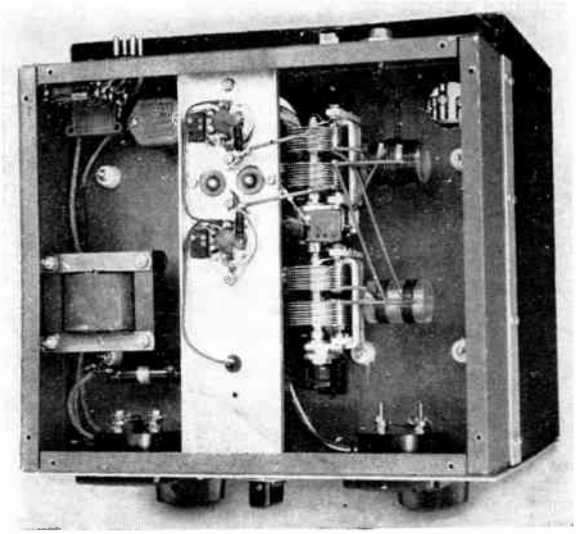

Fig. 6-100 — Bottom view of the push-pull 807 amplifier. The grid-circuit multiband tuner is to the right of the aluminum strip on which the 807 sockets and tubular condensers are mounted. The filament transformer is to the left with the v.h.f. power-lead filters above.

A 1-Kw. Beam-Tetrode Amplifier

Figs. 6-101 through 6-105 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. 6-103. 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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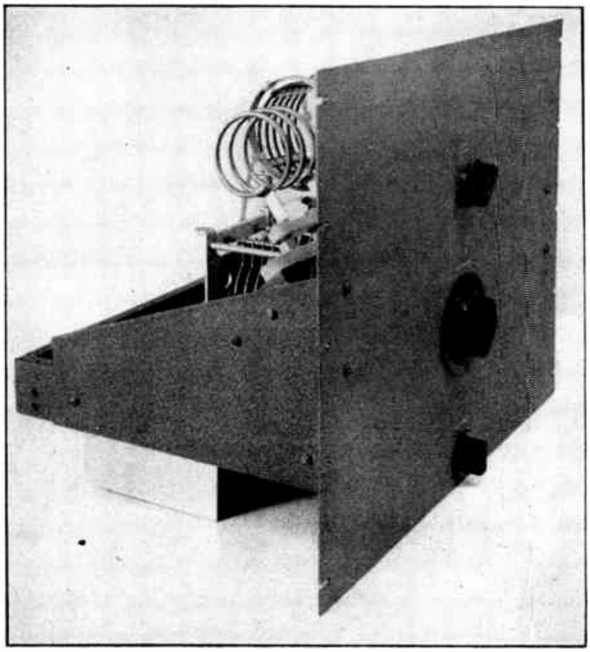

Fig. 6-101 — Front view of the 4-250A amplifier, showing the method of assembling the panel and the chassis. The controls on the panel, from top to bottom, are the output coupling knob, plate tuning dial, and grid tuning dial. The panel is 19 by 17½ inches.

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. 6-104. 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.

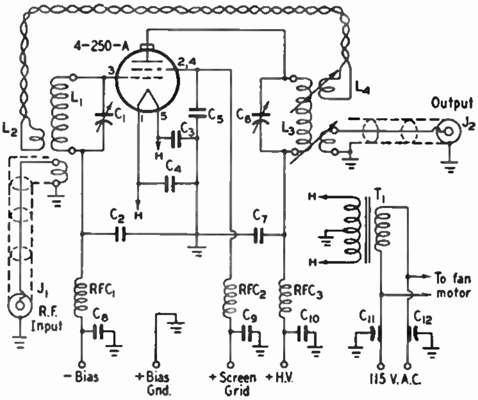

Fig. 6-103 — Schematic diagram of the 4-250A amplifier.

C_1 — 100-$\mu\mu$fd. 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-$\mu\mu$fd. 6000-volt peak (National TMA-150-A).
C_7 — 0.001-μfd. 5000-volt-wkg. mica.
C_8 — 470-$\mu\mu$fd. mica.
C_9 — 500-$\mu\mu$fd. 1000-volt mica.
C_{10} — 500-$\mu\mu$fd. 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).

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Fig. 6-102 — 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.

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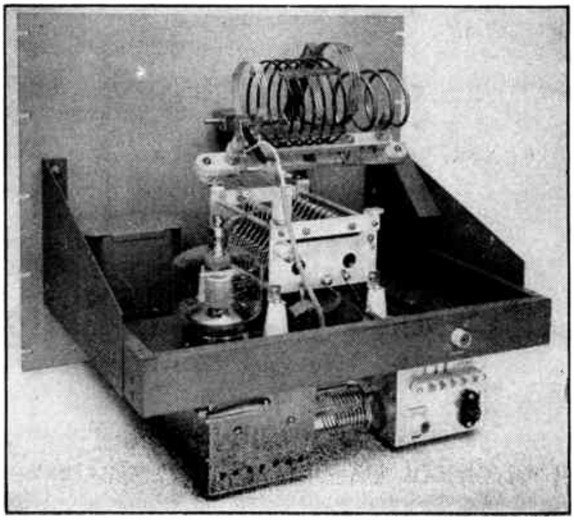

It is advisable to conduct all preliminary adjustments at reduced screen voltage to keep the screen dissipation at a safe level. The lamps, I_1 and I_2 in Fig. 6-104, are for this purpose. A size of lamp should be selected that will give the desired reduction in screen and plate voltage, remembering 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. 6-104, 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.

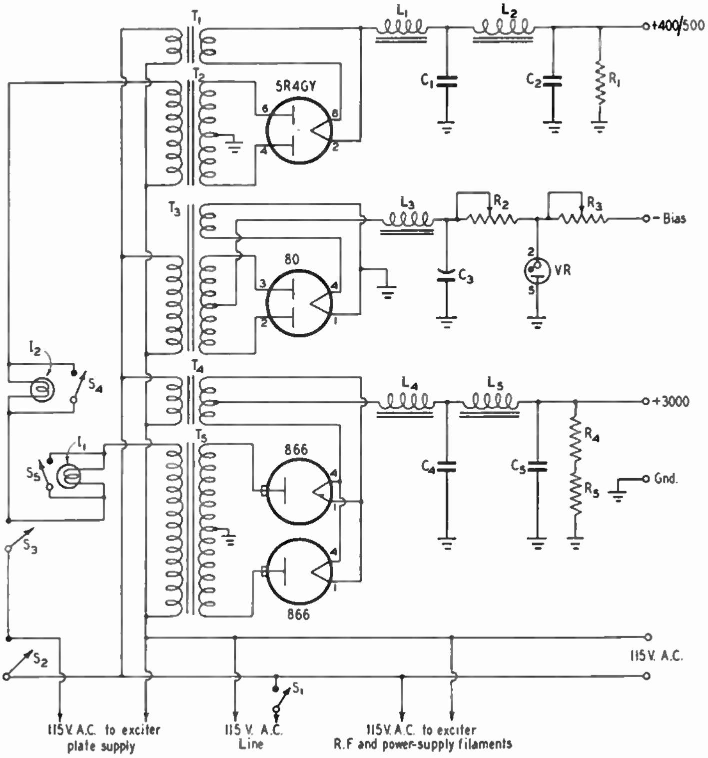

Fig. 6-104 — Circuit diagram of a power supply for the beam-tetrode amplifier. S_1 is the main switch, turning on all filaments. S_2 turns on the plate voltage for the exciter unit and sets up the circuit for S_3 which turns on both screen and plate supplies for the amplifier. I_1 and I_2 are 115-volt 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 — 4-μfd. 600-volt oil-filled.
C_3 — 4-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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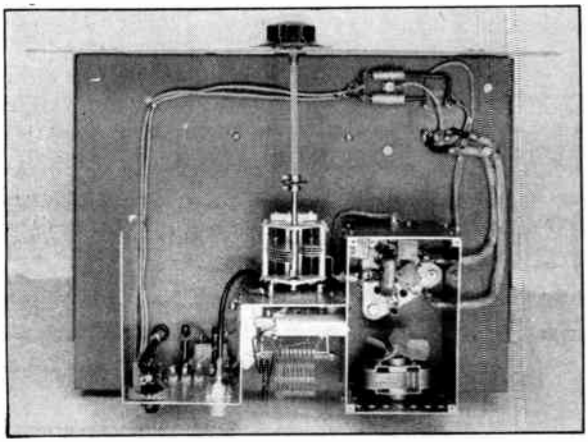

Fig. 6-105 — Bottom view of the 4-250A amplifier. The harmonic filters are in the compartment to the left of the grid coil. The arrangement of the by-pass condensers inside the fan housing is also shown, with the grid terminal of the tube socket pointing toward the grid tank circuit. The chassis measures 17 by 13 by 2 inches.

Rack Construction

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.

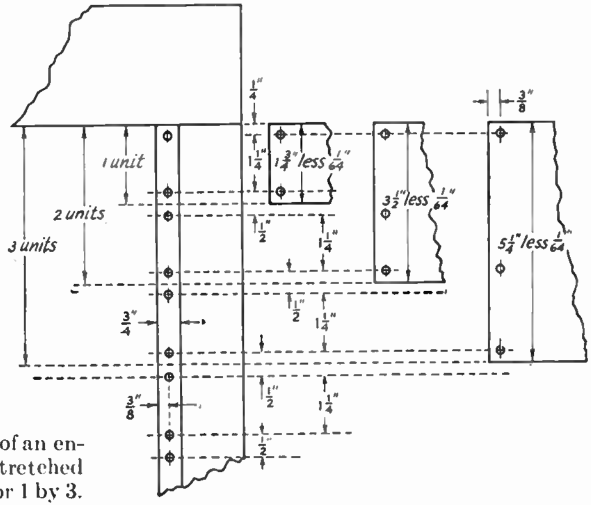

Fig. 6-107 — Detail sketch showing proper drilling for standard rack and panels. As shown for the 3½- and 5¼-inch panels, only sufficient holes are drilled in the panel to provide the necessary strength. When the panels are drilled as shown, they may be moved up and down in steps of 1¾ inches and the holes will always match.

Fig. 6-106A 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. 6-106B, the panel clearance should be 19⅟16 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. 6-107. The table shows the standard drilling for panels of various sizes.

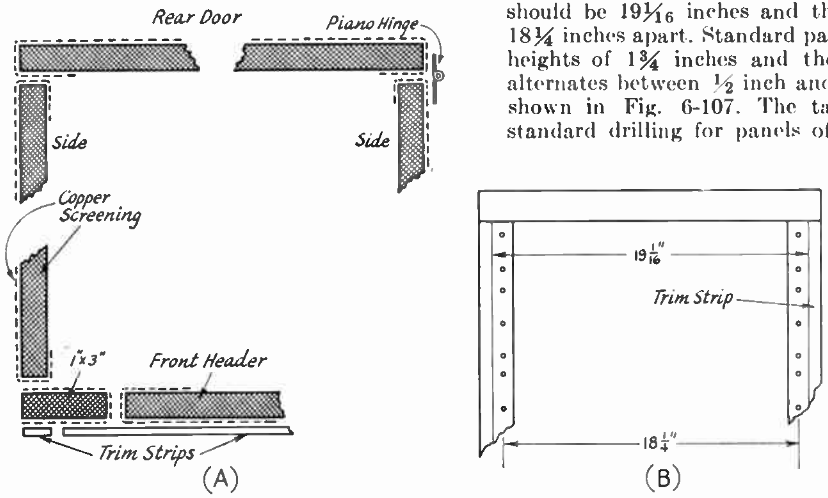

Fig. 6-106 — A — Top detail view of an enclosed relay rack made of wood strips and copper screening. B — Panel-mounting dimensions.

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.

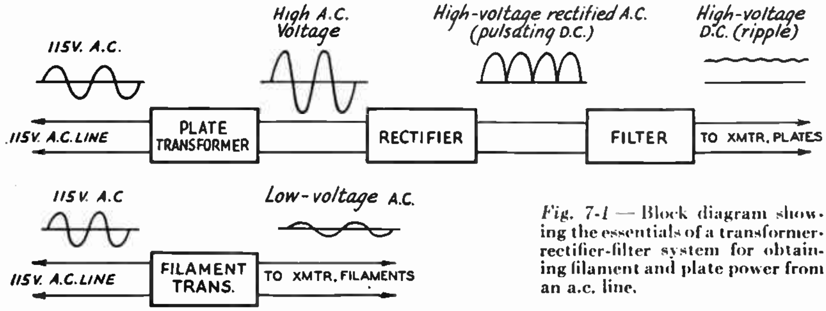

Fig. 7-1 — Block diagram showing the essentials of a transformer-rectifier-filter system for obtaining filament and plate power from an a.c. line.

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.

Another disadvantage of the half-wave rectifier circuit is that the transformer must have a considerably higher primary volt-ampere rating (approximately 40 per cent greater) than in other rectifier circuits.

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

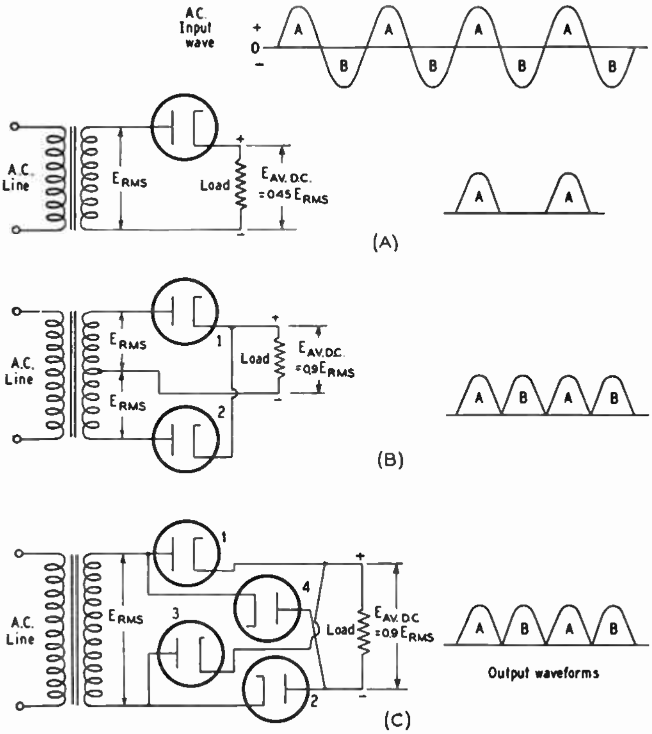

Fig. 7-2 — **Fundamental vacuum-tube rectifier circuits. A — Half-wave. B — Full-wave. C — Bridge.**

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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 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.

When two separate transformers are used in the full-wave circuit with their secondaries connected in series, the same derating mentioned in regard to the half-wave rectifier circuit must be observed.

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 is a diode which requires no cathode heating. Certain types will handle up to 350 ma. at 200 volts d.c. output. The internal 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.

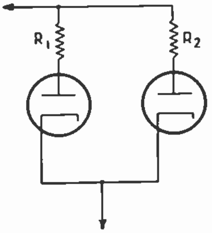

Fig. 7-3 — Connecting rectifiers in parallel for heavier currents. R_1 and R_2 should have the same value, between 50 and 100 ohms.

As with high-vacuum rectifiers, full-wave types are available in the lower-power ratings only. For higher power, two tubes are required in a full-wave circuit.

Selenium Rectifiers

Selenium rectifiers are 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. 7-21 through 7-23. Since they develop little heat if operated within their ratings, they are especially suitable for use in equipment requiring minimum temperature variation.

Typical ratings are listed in the table in the data chapter.

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 and Transformers

The pulsating d.c. wave shown in Fig. 7-2 is not sufficiently smooth to prevent modulation hum. 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 prevent objectionable ripple hum.

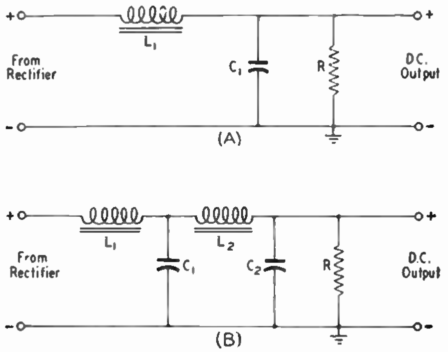

Fig. 7-4 — Choke-input filter circuits. A — Single-section. B — Double-section.

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.

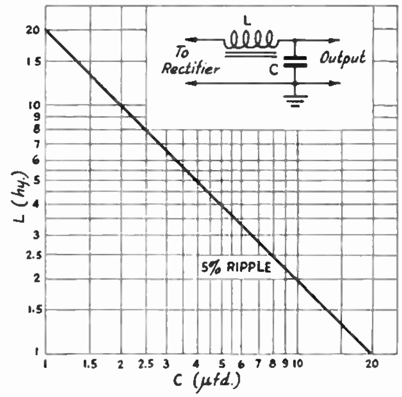

Fig. 7-5 — Graph showing combinations of inductance and capacitance that may be used to reduce ripple to 5 per cent with a single-section choke-input filter.

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:

$$\left.\begin{matrix}\text{Single-}\\\text{Section}\\\text{Filter}\end{matrix}\right\}\ \textit{Percentage ripple} = \frac{100}{LC}$$

where L is in h. and C in μfd.

Example: $L = 5$ h., $C = 4$ 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-h. choke, what capacitance is required to reduce the ripple to 5 per cent?

Referring to Fig. 7-5, following the 10-h. 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 h., 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 minimum 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 approximately by:

$$L_{\text{crit.}} = \frac{\textit{Load resistance}\ (\text{ohms})}{1000}.$$

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.

An inductance of twice the critical value is called the **optimum** value. With this value, the peak plate current of one tube in a center-tap rectifier will be about 10% higher than the d.c. current taken from the supply. To obtain good voltage regulation and, at the same time permit maximum d.c. current to be drawn from the rectifier without exceeding its peak-current rating, the input choke should have the optimum value at full load and the critical value with bleeder only.

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 or 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 h., 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 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 of inductance and capacitance should be used to ensure against resonance effects.

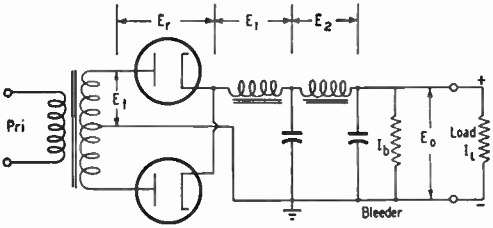

Fig. 7-6 — Diagram showing various voltage drops that must be taken into consideration in determining the required transformer voltage to deliver the desired output voltage.

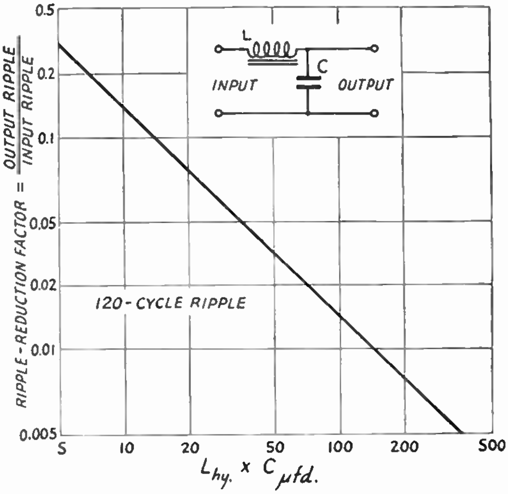

Fig. 7-7 — Ripple-reduction factor for various values of *L* and *C* in filter section. *Output ripple = input ripple × ripple factor.*

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 using the formula previously given.

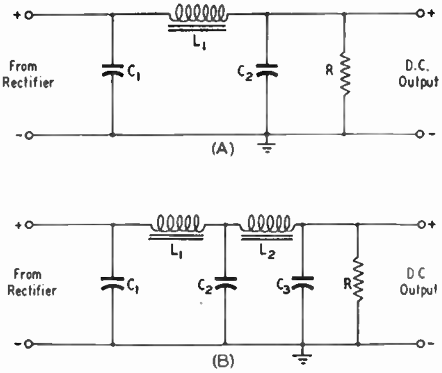

Fig. 7-8 — Condenser-input filter circuits. A — Single-section. B — Double-section.

Additional Filtering

The graph of Fig. 7-7 shows the factor by which the ripple percentage of the first section may be reduced by the addition of one or more sections of filter, each similar in configuration to the first section.

Example:

Ripple after first section = 5 per cent.
L in second section = 10 hy.
C in second section = 8 μfd.
$L \times C$ = 80.

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.

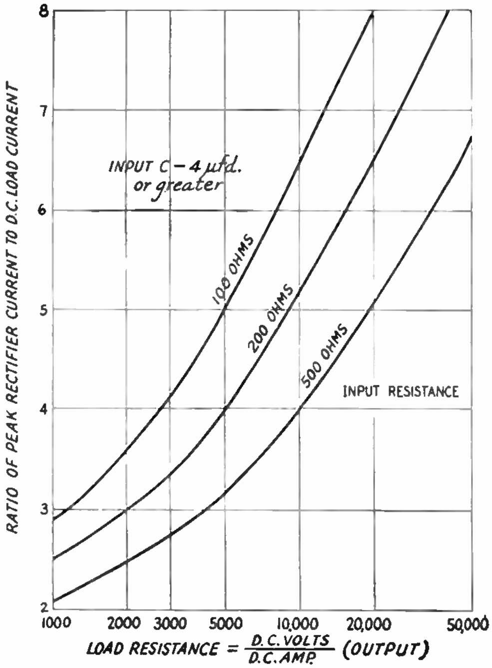

Fig. 7-9 — Graph showing the relationship between the d.c. load current and the rectifier peak plate current with condenser input for various values of load and input resistance.

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.

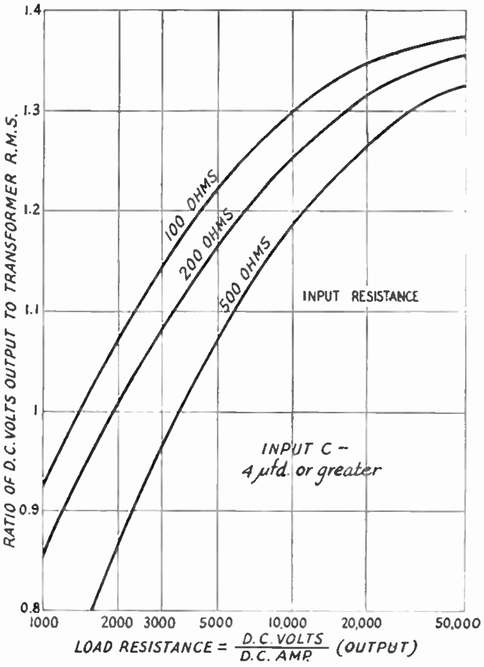

Fig. 7-10 — Chart showing approximate ratio of d.c. output voltage across filter input condenser to transformer r.m.s. secondary voltage for different load and input resistances.

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. 7-10 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. 7-11 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 $= 400 \times 1.17 = 468$ 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 $= 93.6 \times 4 = 374$ 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.

● 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 it would with a filter of the condenser-input type.

In a condenser-input filter, the condensers should have a working-voltage rating at least as high, and preferably somewhat higher, than the peak-voltage rating of the transformer. 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 minimum no-load and full-load inductance values being calculated as described above. For the second choke (**smoothing choke**) 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.

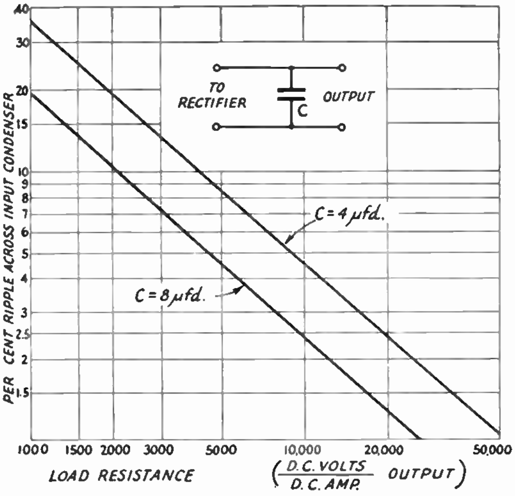

Fig. 7-11 — Chart showing approximate 120-cycle percentage ripple across filter input condenser for various loads.

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 filter circuit.

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 milliamperes, R_1 and R_2 are the d.c. 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.

The approximate transformer output voltage required to give a desired d.c. output voltage with a given load with a condenser-input filter system can be calculated with the help of Fig. 7-10.

Example:

Required d.c. output volts — 500

Load current to be drawn — 100 ma.

Load resistance $= \frac{500}{0.1} =$ 5000 ohms.

If the rectifier resistance is 200 ohms, **Fig. 7-10** shows that the ratio of d.c. volts to the required transformer r.m.s. voltage is approximately 1.15.

The required transformer terminal voltage under load with chokes of 200 and 300 ohms is

$$E_t = \frac{E_o + I\left(\frac{R_1 + R_2}{1000}\right) + E_r}{1.15}$$

$$= \frac{500 + 100\left(\frac{200 + 300}{1000}\right) + 200}{1.15}$$

$$= \frac{570}{1.15} = 495 \text{ volts.}$$

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.

Rewinding Filament Transformers

Although the home winding of high-voltage transformers is a task that few amateurs undertake these days, the rewinding of a small-transformer secondary to give some desired filament voltage is not difficult. It involves a matter of only a small number of turns and the wire is large enough to be handled easily. Often a broadcast-receiver power transformer with a burned-out high-voltage winding, but with the primary winding intact, can be converted into an entirely satisfactory filament transformer without great effort.

The primary volt-ampere rating of a transformer to be rewound may be taken from the label on the transformer or from the manufacturer's catalogue. This will indicate whether or not the transformer will be capable of handling the necessary power. The secondary volt-ampere rating will be ten to twenty per cent less than the primary rating. The product of the voltage and the number of amperes required from the new filament winding, plus that for any other secondaries that may be kept in use, should not exceed the secondary volt-ampere rating, unless the builder is willing to accept a lower safety factor.

Before disconnecting the winding leads from their terminals, each should be marked for identification. In removing the core laminations, care should be taken to note the manner in which the core is assembled, so that the reassembling will be done in the same manner. Some transformers have secondaries wound over the primary, while in others the order is reversed. In case the secondaries are on the inside, the turns can be pulled out from the center after slitting and removing the fiber core.

The turns removed from one of the original filament windings of known voltage should be carefully counted as the winding is removed. This will give the number of turns per volt and the same figure should be used in determining the number of turns for the new secondary. For instance, if the old filament winding was rated at 5 volts and has 20 turns, this is 20/5 = 4 turns per volt. If the new secondary is to deliver 7.5 volts, the required number of turns on the new winding will be 7.5 × 4 = 30 turns.

The *Copper-Wire Table* in the chapter of miscellaneous data shows the current-carrying capacity of various sizes of wire at a cross section of 1500 circular mils per ampere. This is a conservative rating. A cross section of 1000 circular mils per ampere is closer to the figure used for most amateur-service transformers. In cheaper broadcast-receiver transformers, the figure may run as low as 500. The current-carrying capacity at 1000 circular mils per ampere may be determined by pointing off three decimal places from the right in the figures in the third column of the table showing circular-mil area. As an example, No. 18 wire has a capacity of 1.7 amperes at 1500 circular mils per ampere, 2.58 amperes at 1000 circular mils per ampere and 5.16 amperes at 500 circular mils per ampere. The choice of rating to be used in most cases will be decided by the size of available wire and the available winding space. If the transformer being rewound is a filament transformer, it may be necessary to choose the wire size carefully to fit the small available space. On the other hand, if the transformer is a power unit, with the high-voltage winding removed, there should be plenty of room for a size of wire that will conservatively handle the required current.

The insulation to be used between the primary and secondary windings (and also between the secondary winding and the core if the secondary is on the inside) will depend on whether the transformer is to be used to supply r.f. tubes or rectifier tubes in a high-voltage supply. A few layers of linen paper should be sufficient for the former service, but insulating cambric sheet should be used if the voltage between primary and secondary runs more than 1000 volts.

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. 7-12A. 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.

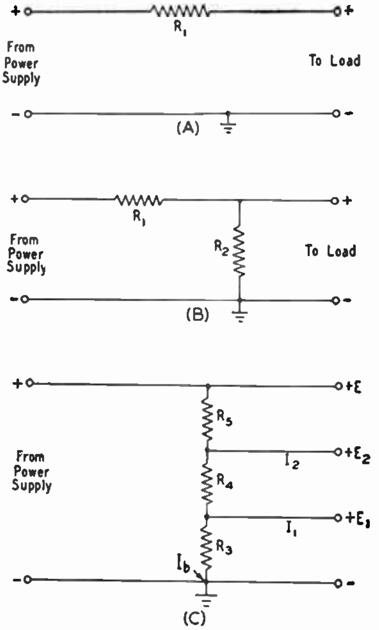

Fig. 7-12 — A — Series voltage-dropping resistor. B — Simple voltage divider. C — Multiple divider circuit.

$$R_3 = \frac{E_1}{I_b};\ R_4 = \frac{E_2 - E_1}{I_b + I_1};\ R_5 = \frac{E - E_2}{I_b + I_1 + I_2}$$

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. 7-12B. 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. 7-12C. 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. 7-12C, I being in decimal parts of an ampere.

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. 7-13A. 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:

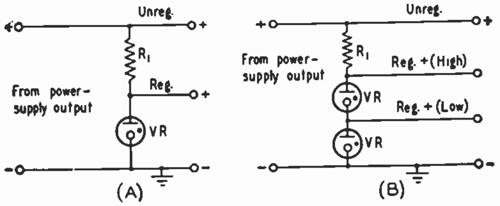

Fig. 7-13 — Voltage-stabilizing circuits using VR tubes.

$$R = \frac{1000\ (E_s - E_r)}{I}$$

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. 7-13B 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 handling 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. 7-14. 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 voltages.

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

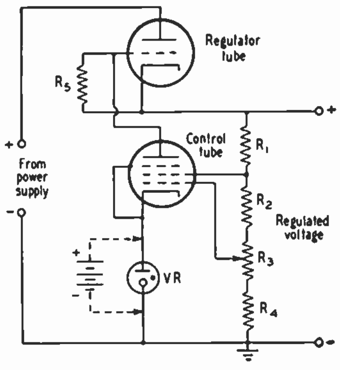

Fig. 7-14 — Electronic voltage regulator. The regulator tube is ordinarily a 2A3 or a number of them in parallel, the control tube a 6SJ7 or similar type. The filament transformer for the regulator tube must be insulated for the plate voltage, and cannot supply current to other tubes when a filament-type 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.

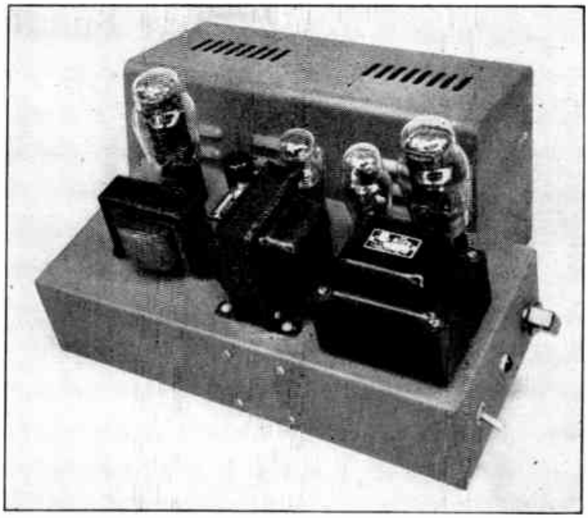

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Fig. 7-15 — A heavy-duty electronically-regulated power supply. The unit is assembled on a 6 × 14 × 3-inch chassis fitted with an enclosing cover. The five tubes across the rear, left to right, are the 6AS7G regulator, the 6SJ7 control tube, the VR-105 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. 7-15. The circuit is shown in Fig. 7-16. 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 (or 6W4GT) 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 300 volts. Filament voltage and an external connection from the bias supply are also brought out to the output socket.

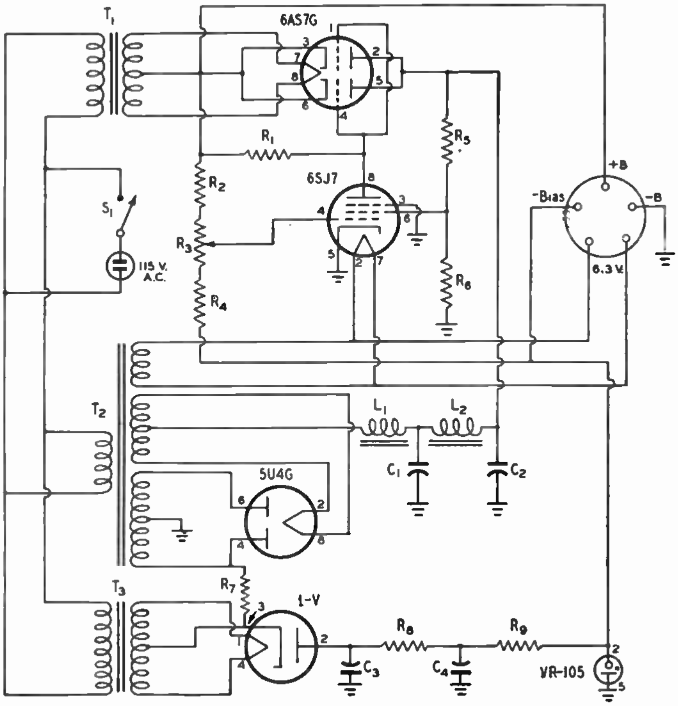

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Fig. 7-16 — Circuit diagram of the electronically-regulated power supply.

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 — 3.5/13-hy. 150-ma. filter choke (Stancor C1718).
L_2 — 7-hy. 150-ma. filter choke (Stancor C1710).
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).

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Bias Supplies

As discussed in the chapter on high-frequency transmitters, 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. 7-17A 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 with rated grid current.

This soaring can be reduced to a considerable extent by the use of a voltage divider across

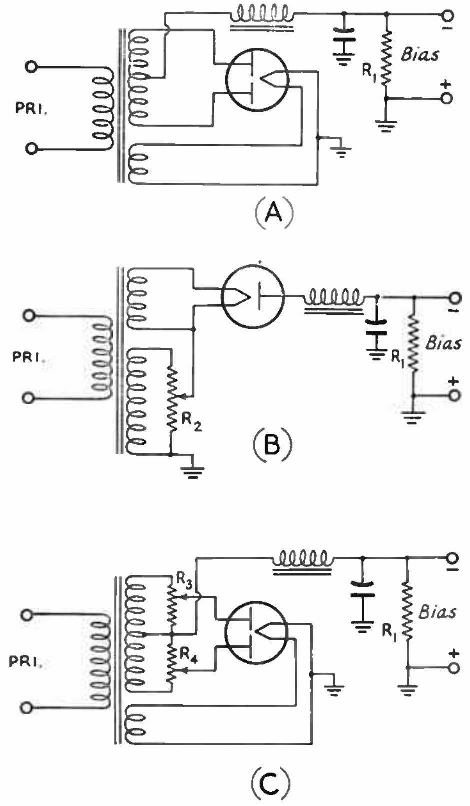


Fig. 7-17 — Simple bias-supply circuits. In A, the peak transformer voltage must not exceed the operating value of bias. The circuits of B (half-wave) and C (full-wave) may be used to reduce transformer voltage to the rectifier. R_1 is the recommended grid-leak resistance.

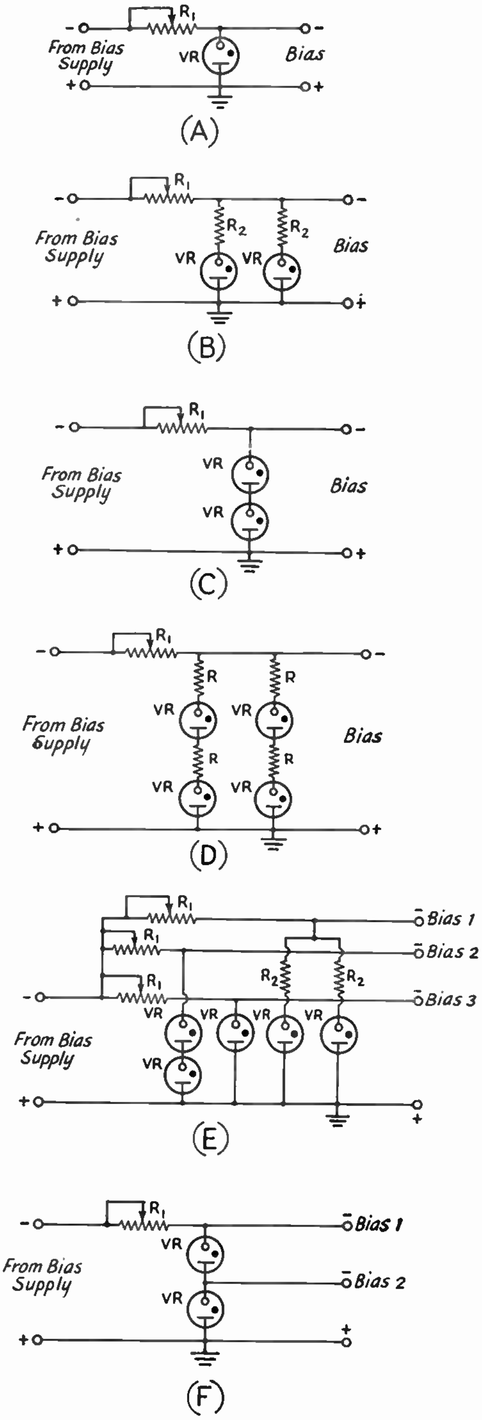


Fig. 7-18 — Illustrating the use of VR tubes in stabilizing protective-bias supplies. R_1 is a resistor whose value is adjusted to limit the current through each VR tube to 5 ma. before amplifier excitation is applied. R and R_2 are current-equalizing resistors of 50 to 100 ohms.

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. 7-17C. 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.

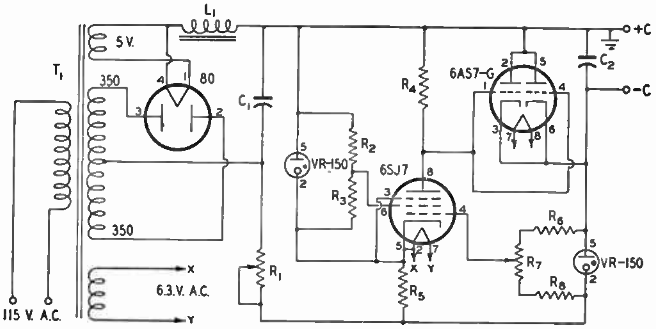


Fig. 7-19 — Circuit diagram of an electronically-regulated bias supply.

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_6 — 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.

Regulated Bias Supplies

The inconvenience of the circuits shown in Fig. 7-17 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. 7-18A. 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 (see transmitter chapter).

Each VR tube will handle 40 ma. of grid current. If the grid current exceeds this value under any condition, similar VR tubes should be added in parallel, as shown in Fig. 7-18B, 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. 7-18C 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. 7-18E, 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. 7-19. 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.

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. 7-20A. In another arrangement, shown at B, a spare filament winding can be used to operate a filament

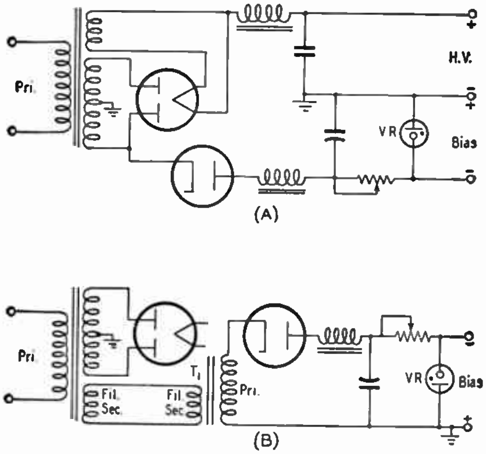


Fig. 7-20 — Convenient means of obtaining biasing voltage. A — From a low-voltage plate supply. B — From spare filament winding. T_1 is a filament transformer, of a voltage output similar to that of the spare filament winding, connected in reverse to give 115 volts r.m.s. output. If cold-cathode or selenium rectifiers are used, no additional filament supply is required.

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.

Selenium-Rectifier Circuits

While the circuits shown in Figs. 7-21, 7-22 and 7-23 may be used with any type of rectifier, they find their greatest advantage when used with selenium rectifiers which require no filament transformer.

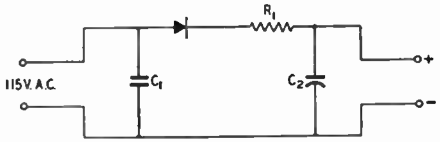


Fig. 7-21 **— Simple half-wave circuit for selenium rectifier.**
C_1 — 0.05-μfd. 600-volt paper.
C_2 — 40-μfd. 200-volt electrolytic.
R_1 — 25 to 100 ohms.

Fig. 7-21 is a straightforward half-wave rectifier circuit which may be used in applications where 115 to 130 volts d.c. is desired. It can be used for 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. 7-18 shows several voltage-doubler circuits. Of the three, the one shown at A 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 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.

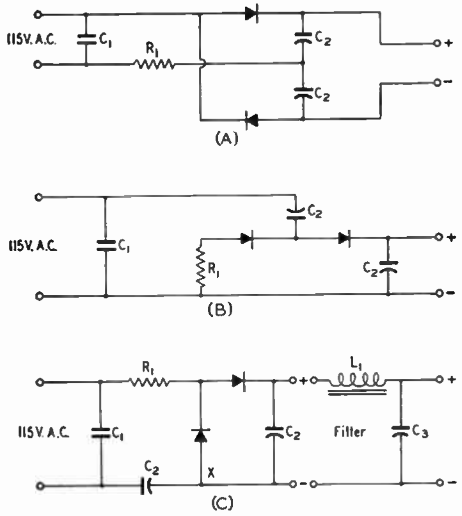


Fig. 7-22 **— Voltage-doubling circuits for use with selenium rectifiers.**
C_1 — 0.05-μfd. 600-volt paper.
C_2 — 40-μfd 200-volt electrolytic.
C_3 — Filter condenser.
R_1 — 25 to 100 ohms.
L_1 — Filter choke.

Fig. 7-23A and B shows voltage-tripler and voltage-quadrupler circuits respectively, for use where higher voltages are desired. They can be used for powering the small transmitter.

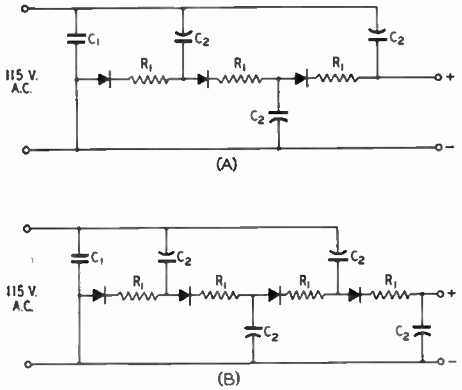


Fig. 7-23 **— Selenium-rectifier voltage-tripling and voltage-quadrupling circuits.**
C_1 — 0.05-μfd. 600-volt paper.
C_2 — 40-μfd. 450-volt electrolytic.
R_1 — 25 to 100 ohms.

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. 7-23 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. 7-22C, will provide sufficient smoothing for most applications.

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.

● POWER-LINE CONNECTIONS

If the transmitter is rated at much more than 100 watts, special consideration should be given to the a.c. line running into the station. In some residential systems, three wires are brought in from the outside to the distribution board, while in other systems there are only two wires. In the three-wire system, the third wire is the **neutral** which is grounded. The voltage between the other two wires normally is 230, while half of this voltage (115) appears between each of these wires and neutral, as indicated in Fig. 7-24A. In systems of this type, usually it will be found that the 115-volt household load is divided as evenly as possible between the two sides of the circuit, half of the load being connected between one wire and the neutral, while the other half of the load is connected between the other wire and neutral. Heavy appliances, such as electric stoves and heaters, normally are designed for 230-volt operation and therefore are connected across the two ungrounded wires. While both ungrounded wires should be fused, a fuse should never be used in the wire to the neutral, nor should a switch be used in this side of the line. The reason for this is that opening the neutral wire does not disconnect the equipment. It simply leaves the equipment on one side of the 230-volt circuit in series with whatever load may be across the other side of the circuit, as shown in Fig. 7-24B. Furthermore, with the neutral open, the voltage will then be divided between the two sides in proportion to the load resistance, the voltage on one side dropping below normal, while it soars on the other side, unless the loads happen to be equal.

The usual line running to baseboard outlets is rated at 15 amperes. Considering the power consumed by filaments, lamps, modulator, receiver and other auxiliary equipment, it is not unusual to find this 15-ampere rating exceeded by the requirements of a station of only moderate power. It must also be kept in mind that the same branch may be in use for other household purposes through another outlet. For this reason, and to minimize light blinking when keying or modulating the transmitter, a separate heavier line should be run from the distribution board to the station whenever possible. (A three-volt drop in line voltage when the load is applied will cause noticeable light blinking.)

If the system is of the three-wire type, the three wires should be brought into the station so that the station load can be distributed to keep the line as balanced as possible. The voltage across a fixed load on one side of the circuit will increase as the load current on the other side is increased. The rate of increase will depend upon the resistance introduced by the neutral wire. If the resistance of the neutral is low, the increase will be correspondingly

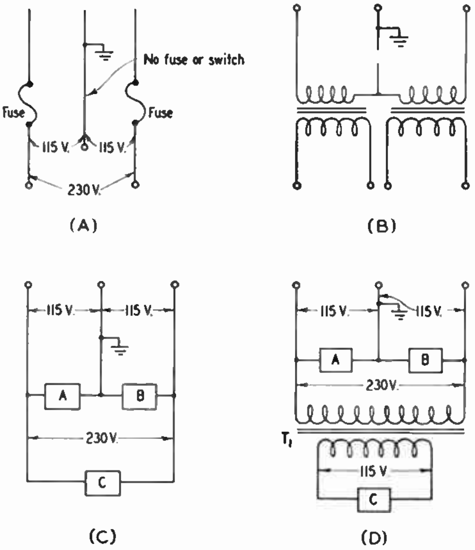


Fig. 7-24 — Three-wire power-line circuits. A — Normal 3-wire-line termination. No fuse should be used in the grounded (neutral) line. B — Showing that a switch in the neutral does not remove voltage from either side of the line. C — Connections for both 115- and 230-volt transformers. D — Operating a 115-volt plate transformer from the 230-volt line to avoid light blinking. T_1 is a 2-to-1 step-down transformer.

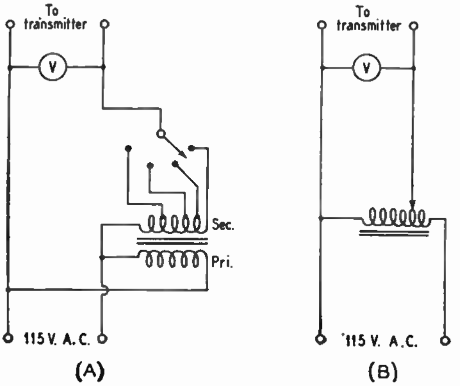
small. When the currents in the two circuits are balanced, no current flows in the neutral wire and the system is operating at maximum efficiency.

Light blinking can be minimized by using transformers with 230-volt primaries in the power supplies for the keyed or intermittent part of the load, connecting them across the two ungrounded wires with no connection to the neutral, as shown in Fig. 7-24C. The same can be accomplished by the insertion of a step-down transformer whose primary operates at 230 volts and whose secondary delivers 115 volts. Conventional 115-volt transformers may be operated from the secondary of the step-down transformer (see Fig. 7-24D).

When a special heavy-duty line is to be installed, the local power company should be consulted as to local requirements. In some localities it is necessary to have such a job done by a licensed electrician, and there may be special requirements to be met in regard to fittings and the manner of installation. Some amateurs terminate the special line to the station at a switch box, while others may use electric-stove receptacles as the termination. The power is then distributed around the station by means of conventional outlets at convenient points. All circuits should be properly fused.

● 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 at evening, they may be taken care of by the use of a manually-operated compensating device. A simple arrangement is shown in Fig. 7-25A. 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.



(A) (B)

Fig. 7-25 — 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) which feeds the transformer primaries.

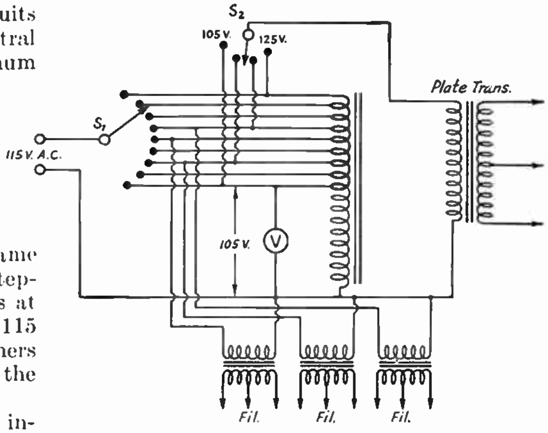


Fig. 7-26— With this circuit, a single adjustment of the tap switch S_1 places the correct primary voltage on all transformers in the transmitter. Information on constructing a suitable autotransformer at negligible cost is contained in the text. The light winding represents the regular primary winding of a revamped transformer, the heavy winding the voltage-adjusting section.

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 7-25B illustrates the use of a variable transformer (Variac) for adjusting line voltage to the desired value.

Another scheme by which the primary voltage 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. 7-26.

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

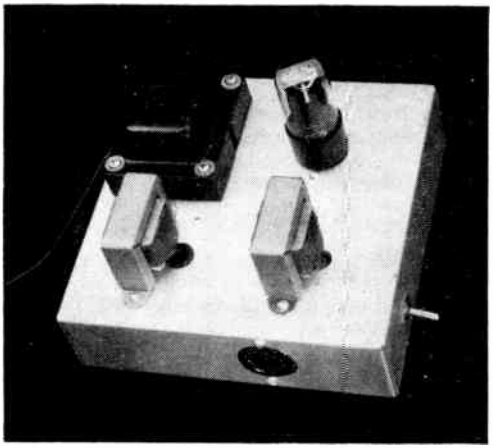


Fig. 7-27— A typical simple receiver power supply. Filament and plate voltages are taken from the multi-contact tube socket which serves as an outlet.

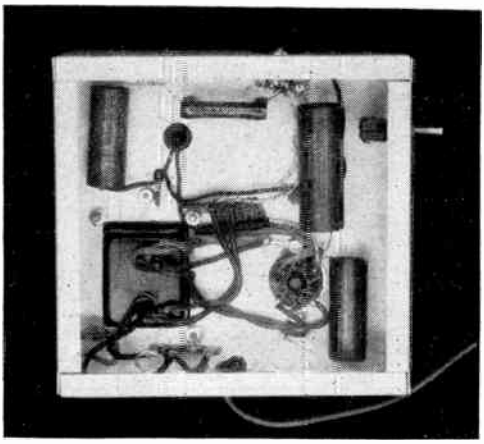


Fig. 7-28— Bottom view of the receiver power supply showing the cut-out for the flush-mounting transformer.

permits power-input control and does not require an extra autotransformer.

Constant-Voltage Transformers

Although comparatively expensive, special transformers called **constant-voltage transformers** are available for use in cases where it is necessary to hold line voltage and/or filament voltage constant with fluctuating supply-line voltage. They are rated over a range of 17 va. at 6.3 volts output, for small tube-heater demands, up to several thousand volt-amperes at 115 or 230 volts. In average figures, such transformers will hold their output voltages within one per cent under an input-voltage variation of 30 per cent.

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. Power-supply units should be fused individually.

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

◆

Fig. 7-29 — A typical high-voltage transmitter power supply. The transformers, chokes and condensers are inverted so that no terminals are exposed to accidental contact. The caps of the 866 rectifiers are the insulated type.

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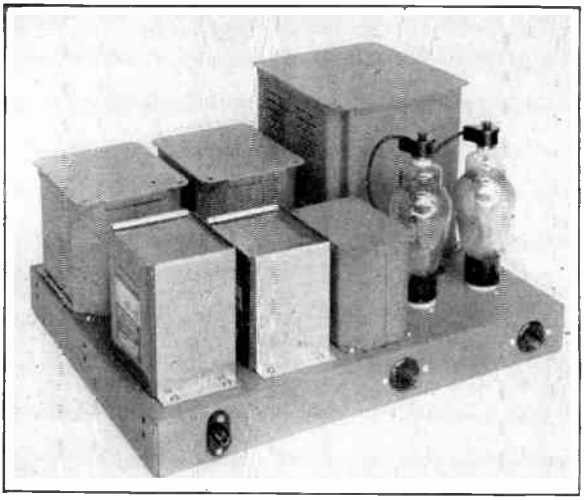

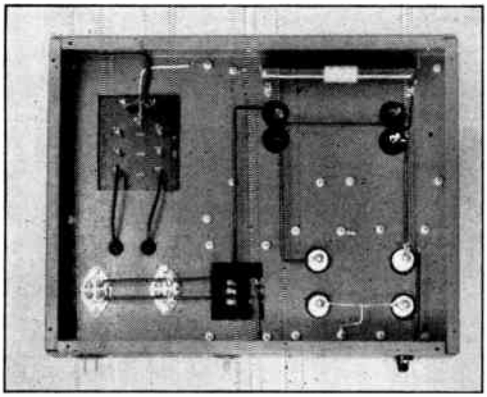

Fig. 7-30 — **Bottom view of the transmitter power supply showing the cut-outs for the terminals. Separate power plugs are used for the rectifier-filament and plate transformers so that they may be switched independently from the control position.**

voltage-dropping resistors should be placed where they are open to air circulation. Placing them in confined space reduces the 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 filament and plate 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 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. 7-31 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.

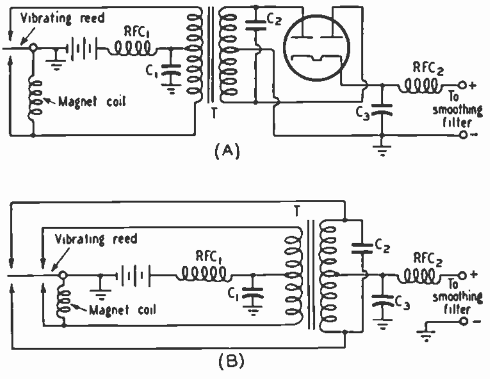

Fig. 7-31 — Basic types of vibrator power-supply circuits. A—Nonsynchronous. B—Synchronous.

The synchronous circuit of Fig. 7-31B 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.

Vibrator-transformer units are available in a variety of power and voltage ratings. Representative units vary from one delivering 125 to 200 volts at 100 ma. to others that have a 400-volt output rating at 150 ma. Most units come supplied with "hash" filters, but not all of them have built-in ripple filters. The requirements for ripple filters are similar to those for a.c. supplies. The usual efficiency of vibrator packs is in the vicinity of 70 per cent, so a 300-volt, 200-ma. unit will draw approximately 15 amperes from a 6-volt storage battery. Special vibrator transformers are also available from transformer manufacturers so that the amateur may build his own supply if he so desires. These have d.c. output ratings varying from 150 volts at 40 ma. to 330 volts at 135 ma.

"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 power supply and its 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 often found to be particularly helpful.

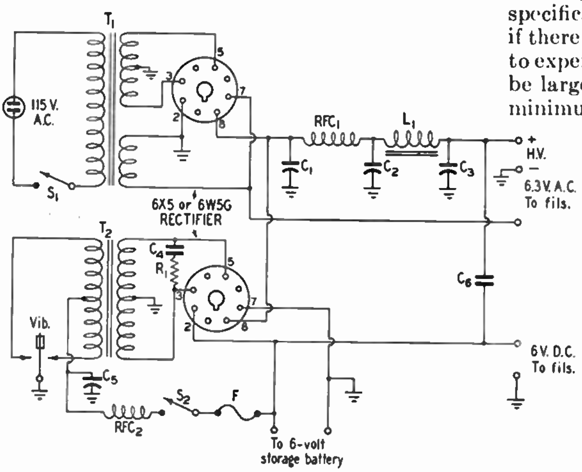

Fig. 7-32 — Circuit of a combination a.c.-d.c. power supply for emergency work.

C_1 — 0.01-µfd. 600-volt paper.
C_2 — 8-µfd. 450-volt electrolytic.
C_3 — 32-µfd. 450-volt electrolytic.
C_4 — 0.005-to 0.01-µfd. 1600-volt paper.
C_5 — 0.5-µfd. paper, 50 volts or higher.
C_6 — 100-µµfd. 600-volt mica.
R_1 — 4700 ohms, 1 watt.
L_1 — 10- to 12-hy. filter choke, 100 ma. (not over 100 ohms) (Stancor C-2303 or equivalent).
RFC_1 — 2.5-mh. r.f. choke.
RFC_2 — 55 turns No. 12 on 1-inch form, close-wound.
S_1, S_2 — Toggle switch.
T_1 — Power transformer: 275 to 300 volts r.m.s. each side of center tap, 100 to 150 ma., 6.3-volt filament winding.
T_2 — Vibrator transformer (Stancor P-6131 or similar).
VIB — Vibrator unit (Mallory 500P, 294, etc.).

PRACTICAL VIBRATOR-SUPPLY CIRCUIT

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.

An example of the latter-type circuit is shown in Fig. 7-32. It consists essentially of two transformer-rectifier systems — one for 115 volts a.c. and the other a vibrator system to operate from a 6-volt storage battery. A common filter is used for the two systems. In interchanging between a.c. and d.c. operation, the rectifier tube (a 6X5 or 6W5G) is shifted to the appropriate socket, while the filament connections are made to the proper output terminals. If desired, two rectifier tubes may be used and the changeover made through suitable switches.

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 components are assembled on a 5 × 10 × 3-inch steel 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.

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. 7-34 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

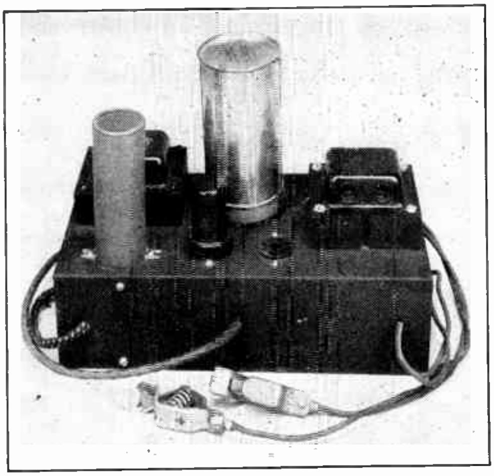

Fig. 7-33 — A typical combination a.c.-d.c. power pack for low-power emergency work. The two transformers are mounted at either end of the chassis. The filter condenser is at the left, the two rectifier sockets at the center and the vibrator to the rear.

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. 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.

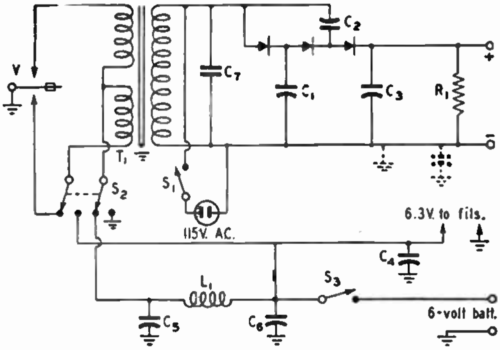

Fig. 7-34 — Circuit diagram of a compact vibrator-a.c. portable power supply using selenium rectifiers.

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.

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 cannot be connected to actual ground except through by-pass condensers.

● 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 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 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.

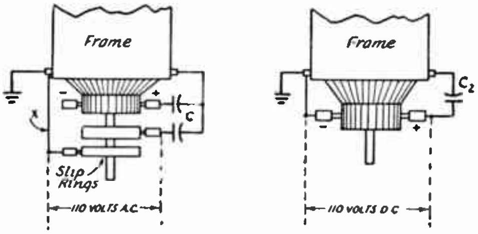

Fig. 7-35 — Connections used for eliminating interference from gas-driven generator plants. *C* should be 1 µfd., 300 volts, paper, while C_2 may be 1 µfd. with a voltage rating of twice the d.c. output voltage delivered by the generator. X indicates an added connection between the slip ring on the grounded side of the line and the generator frame.

From this point on, if necessary, by-pass condensers from various brush holders to the frame, as shown in Fig. 7-35, 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.

Table 7-I gives service life of representative types of B 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.

Dry "B" batteries are made in a variety of sizes and shapes, from a 45-volt unit weighing about 1 lb. that has an intermittent service rating of 20 hours at a drain of 20 ma., to a 12-lb. unit rated at 130 hours at 40 ma. "A" batteries for filament service range from a 6-volt unit weighing 1½ lbs. delivering in intermittent service and average of 60 ma. for 150 hours, to a 6¼-lb. 1.5-volt unit having a service life of 870 hours at 200 ma. Miniature batteries, suitable for hand-portable use, are also available.

TABLE 7-I — 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.	Volts	5	10	15	20	25	30	40	50	60	100	150
	759	17	9	90	Suggested current range = 7 to 12 Ma.										
—	758	14	4	90											
21308	—	12	14	45	1600	1100	690	490	—	300	200	—	130	60	30
10308	—	11	4	45	1300	750	520	350	—	—	130	—	90	45	22
—	754	6	1	90	Suggested current range = 5 to 15 Ma.										
2308	—	8	3	45	1100	500	330	200	—	150	80	—	43	—	—
—	487	4	2	45	1000	430	250	170	—	—	—	—	—	—	—
B30	—	2	—	45	350	170	90	50	—	21	17	—	—	—	—
A30	—	2	—	45	260	100	48	28	—	17	7	—	—	—	—
—	482	1	15	45	400	210	122	80	—	—	—	—	—	—	—
Z30NX	—	1	4	45	155	70	30	20	15	9.5	—	—	—	—	—
—	738	1	4	45	130	52	28	17	—	—	—	—	—	—	—
	490	—	15	90	70	20									
—	467	—	12	67.5	70	20	—	—	—	—	—	—	—	—	—
	457	—	8	67.5	35	9									
—	455	—	8	45	70	20	—	—	—	—	—	—	—	—	—
XX30	—	—	8	45	70	25	12	7	—	3.5	—	—	—	—	—

CHAPTER 8

Keying and Break-In

Offhand it would appear that keying a transmitter is a simple matter, since on the face of it nothing more is involved than turning the transmitter output on and off to correspond to the code characters being sent. Unfortunately, it is not this simple, and perfect keying of a c.w. rig is as difficult to come by as perfect voice quality is with a 'phone transmitter. The problem cannot be dismissed lightly.

Although the operation is basically that of turning the transmitter output power on and off, it is complicated by the fact that it must not be turned on and off *instantaneously*. Instead, the output must be made to rise to (and fall from) maximum in some finite period of time, if **key clicks** are to be avoided. These clicks are the inescapable result of changing the power level rapidly, and they appear in the radio spectrum adjacent to the signal proper. The more rapidly the output is varied, the farther the clicks will extend in frequency and the greater will be their amplitude. They interfere unnecessarily with other signals and, if severe enough, can be cause for a discrepancy report by the FCC.

Another effect of improper keying of a transmitter is the introduction of **chirp**, a change in frequency at the instant of making or breaking the signal. A chirp of 50 cycles is enough to make a signal unpleasant to copy, and a chirp of several hundred cycles may render the signal difficult to copy or a target for an FCC discrepancy report. Much depends, of course, upon the selectivity and beat note being used at the receiver, but the safest procedure is to aim for no detectable chirp.

A third keying fault is defined as **backwave**, and it consists of power leaking through and being radiated when the key is "up." If strong enough, backwave makes the signal unpleasant or difficult to copy.

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 often desirable 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, unless the oscillator stage is well shielded.[1] However, chirpless and clickless keying of an oscillator is difficult to obtain, since the necessary slow turning on and off of the oscillator (for click elimination) shows up any oscillator frequency-vs.-voltage changes. It is easy to key an oscillator without chirps or without clicks but not without both. Since the effect of a chirp is multiplied with frequency, it is quite difficult to obtain chirpless oscillator keying at an output frequency of 14 or 28 Mc.

The best-sounding keying (and the most simple to adjust) is usually obtained by keying the output or driver stage, or both. With the oscillator running continuously and "buffered" by several intermediate stages, its frequency remains constant throughout all parts of the keying cycle. The only problem in keying then becomes that of properly "shaping" the keying to reduce or eliminate clicks. When keying several stages away from the output amplifier, it is necessary to bias the stages following the keyed stage so that they draw little or no plate current when the key is up, to avoid excessive plate dissipation. If the stages are biased too heavily, however, these subsequent amplifiers tend to shorten the rise and fall times and thus reintroduce clicks. This should always be borne in mind when a multistage transmitter is used with oscillator or other low-level keying.

The power broken by the key is an important consideration, both from the standpoint of safety to the operator and that of sparking and sticking at the key contacts. Keying of the oscillator or a low-power stage is favorable on both counts. The use of a keying relay or keyer tube is recommended when a high-power circuit is keyed.

Because transmitters vary widely in design,

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[1] For a description of a well-shielded oscillator, see Smith, "A Solution to the Keyed-VFO Problem," *QST*, February, 1950.

there is no specific recommendation that can be made about choosing the stage to key. If the oscillator alone keys satisfactorily (no chirps or clicks), even when listening to its harmonics on 14 or 28 Mc., the transmitter should be keyed there, but the effect of adding the additional multipliers and amplifiers should be carefully checked, to see that clicks are not reintroduced. Methods for checking will be given later. If the oscillator cannot be keyed satisfactorily by itself or with the following stage added, a stage near the output should be keyed and any thought of break-in operation should be discarded. A close approach to break-in operation can be obtained by using a convenient and fast "on-off" switch for the oscillator, or the break-in system described later in the chapter can be used.

Keying Circuits

The plate circuit is a good one to key in an oscillator or low-voltage amplifier, because it is easy to shape the keying properly in this circuit. When **plate-circuit keying** is used, however, it is usually done in the negative lead, since this permits one side of the key to be grounded. The stage can be keyed in the positive lead, but both sides of the keyed circuit will be "hot," and a keying relay is advisable. Fig. 8-1 shows the general circuit for negative-lead keying in either an oscillator or an amplifier. Two examples are shown using triodes, but screen-grid tubes can be used just as readily. Plate-circuit keying is recommended only for low-voltage circuits if no keying relay is used, since a large portion of the supply voltage can appear across the open key.

Shaping circuits applicable to this and later circuits will be discussed in this chapter under "Testing Your Keying."

Somewhat closely related to plate-circuit keying is **screen-grid keying,** shown in Fig. 8-2. The only basic difference is that the screen grid is pulled down to a negative voltage when the key is up, to avoid the backwave that may be present when the screen goes only to zero volts. The negative supply can be small, since its current demand is only a few milliamperes. If the screen voltage is taken from the plate supply, it should come from a voltage divider rather than a simple dropping resistor.

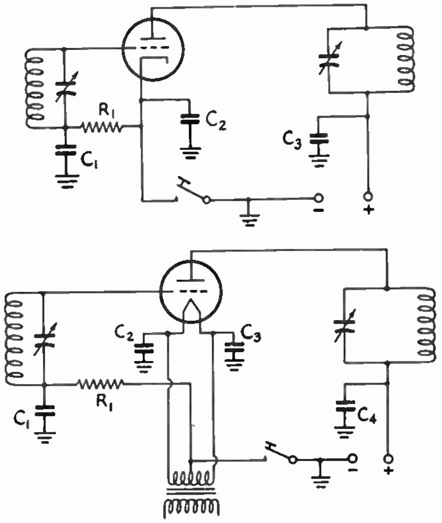

Fig. 8-1 — Negative plate-lead keying for cathode- or filament-type tubes. These circuits are useful for oscillator or low-power stages, where the voltage across the open key is not very dangerous. Tetrode or pentode stages can be keyed in this manner, but the screen circuit should be stabilized with VR tubes or a heavy voltage divider. R_1 is the normal grid leak, C_1, C_2, C_3 and C_4 are r.f. by-pass condensers.

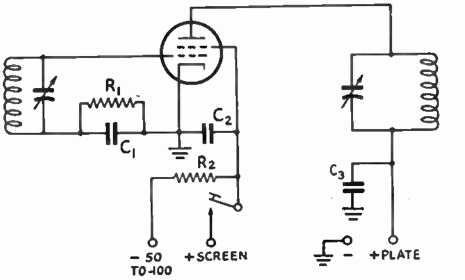

Fig. 8-2 — Screen-grid keying, suitable for oscillator or amplifier keying. R_1 is the normal grid leak, R_2 should be about 200 to 500 ohms per screen volt, and C_1, C_2 and C_3 are normal by-pass condensers.

Grid-circuit, or **blocked-grid,** keying is shown in Fig. 8-3. With the key up, a negative voltage is applied to the grid sufficient to cut off the tube and prevent current flow. With the key closed, the grid circuit develops normal grid bias through R_2. The drain on the negative-voltage supply is small, since it is limited by the size of R_1. Grid-circuit keying is most generally used with low-power stages or where the voltage necessary to cut off the amplifier is only a few hundred volts. The value of C determines the keying characteristic, together with the ratio of R_2 and R_1, and will be discussed later.

By placing the key in the cathode (or center tap) circuit of an oscillator or amplifier, both the grid and plate (and screen, if any) circuits are opened by the key. **Cathode keying** is good for use with amplifiers, because the proper

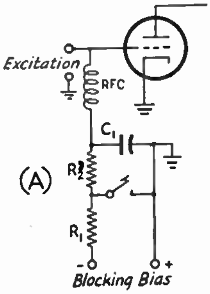

Fig. 8-3 — Blocked-grid keying. R_1, the current-limiting resistor, should have a value of about 50,000 ohms. C_1 may have a capacity of 0.1 to 1 μfd., depending upon the keying characteristic desired. R_2 is the normal value of grid leak for the tube.

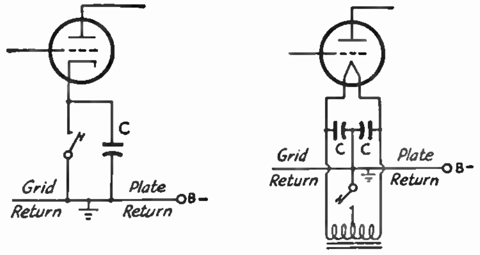

Fig. 8-4 — Cathode and center-tap keying. The condensers *C* are r.f. by-pass condensers. Their capacity is not critical, values of 0.001 to 0.01 μfd. ordinarily being used.

shaping can be accomplished readily. It is also widely used with oscillators, but here the shaping is often complicated by the grid-circuit time constant. Cathode keying is shown in Fig. 8-4. It is popular for use in low- and medium-power stages, although a keying relay or keyer tube should be used where the plate voltage is more than 300.

A popular method of keying involves using one or more tubes as **keyer tubes**, in place of a relay. A keyer tube (or tubes) can be used in the negative-lead or cathode-keying circuits of Figs. 8-1 and 8-4. One advantage of tube keying is that the voltage across the key is limited by large resistors, and so the operator has no chance for anything but the slightest electrical shock. A further advantage is that the shaping is done in the grid circuit of the keyer tube with inexpensive parts. The basic keyer tube circuit is shown in Fig. 8-5 — it is similar to the grid-circuit keying of Fig. 8-3.

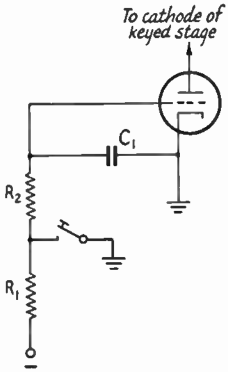

Fig. 8-5 — The basis keyer-tube circuit for cathode or negative-lead keying.

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.

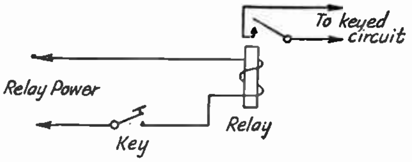

Fig. 8-6 — A keying relay can always be substituted for the key, to provide better isolation from the keyed circuit. An r.f. filter is generally required at the key, and the keying filter is connected in the keyed circuit at the relay contacts.

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.

Testing Your Keying

The choice of a keying circuit is not as important as its complete testing. Any of the circuits shown can be made to give satisfactory keying, but they must be adjusted properly.

The easiest way to find out what your keyed signal sounds like on the air is to trade stations with a near-by ham friend some evening for a short QSO. If he is a half mile or so away, that's fine, but any distance up to the point where the signals are still S9 will be satisfactory.

After you have found out how to work his rig, make contact and then have him send slow dashes, with dash spacing. (The letter "T" at about 5 w.p.m.) With the crystal filter out, cut the r.f. gain back just enough to avoid receiver overloading (the condition where you get crisp signals instead of mushy ones) and tune slowly from out of beat-note range on one side of the signal through to zero and out the other side. Knowing the tempo of the dashes, you can readily identify any clicks in the vicinity as yours or someone else's. A good signal will have a thump on "make" that is perceptible only where you can also hear the beat note, and the click on "break" should be practically negligible at any point. Fig. 8-7A shows how it should sound. If your signal is like that, it will sound good, provided there are no chirps. Then have him run off a string of 35- or 40-w.p.m. dots with the bug — if they are easy to copy, your signal has no "tails" worth worrying about and is a good one for any speed up to the limit of manual keying. If the receiver has poor selectivity with the crystal filter out, make one last check with the filter in (Fig. 8-7B), to see that the clicks off the signal are negligible even at high signal level.

If you don't have any convenient friends with whom to trade stations, you can still check your keying, although you have to be a little more careful The first step is to get rid of the r.f. click at the key, because if you don't you will never know where you stand. Locally (meaning in your own receiver) this click will coincide in time with clicks that may or may not be on your signal, so there is just no way to observe your signal without first eliminating the r.f. click. And unless you have a keying system that breaks no current, you have a

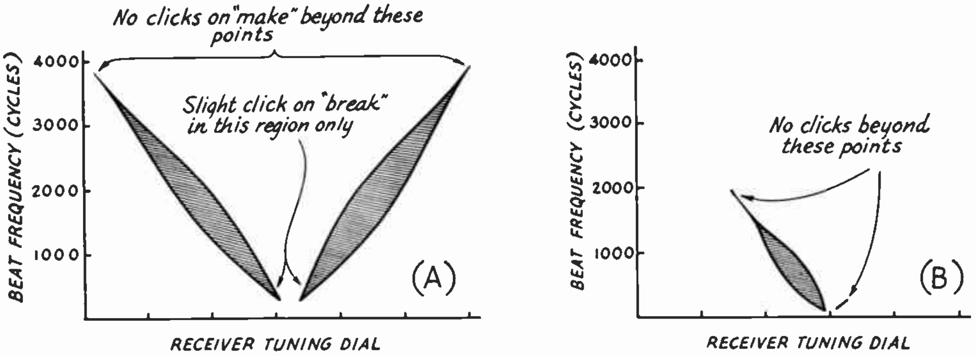

Fig. 8-7 — Representations of a clean c.w. signal as a receiver is tuned through it. (A) shows a receiver with no crystal filter and the b.f.o. set in the center of the passband, and (B) shows the crystal filter in and the receiver adjusted for single-signal reception. The variation in thickness of the lines represents the relative signal intensity. The audio frequency where the signal disappears will depend upon the receiver selectivity characteristic and the strength of the signal.

click at the key. Even the current broken by the key in a vacuum-tube keyer circuit (which is sometimes only 0.1 ma. or so) will cause r.f. clicks that can be heard in your receiver and often in the b.c. set. If you key with a relay, the key opens the relay-coil circuit and clicks are generated at the key as well as at the relay contacts. Don't make the very common mistake of thinking these clicks are the same as the on-the-air clicks discussed earlier — they are not! They are simply local clicks that you must eliminate before you can observe your signal in your receiver. These clicks are the same as the ones you get when you turn an electric light on or off — when you suddenly start or stop current flow, no matter how little, you generate r.f. and that's the click.

Getting rid of this little click is generally no trick at all, unless you're breaking a lot of current. All it requires is a small r.f. filter, as shown in Fig. 8-8. Sometimes just a small (0.001-μfd.) condenser mounted right at the key terminals will do it, and sometimes it will require the full treatment complete with r.f. chokes and second condenser. Measure the normal current through the key leads, remove the transmitter leads, and then connect a d.c. power supply and resistor to give the same current through the key. When your key will break this current with no click, as observed in your receiver and the b.c. set (tuned off any station), you have a suitable r.f. filter at the key and you can reconnect the transmitter. If you use a vacuum-tube keyer, just don't turn on the transmitter but key the normal keyer grid current. If you use a keying relay, first eliminate the click at the key by just keying the relay and adding filter across the key, and then eliminate the click at the relay contacts with another r.f. filter in the relay-keyed circuit. The filter should be mounted right at the key or relay contacts. The objective is to be able to make or break normal key current without generating a local click, and the filtering is usually so simple that the junk box will yield the parts and the process takes longer to describe than to apply.

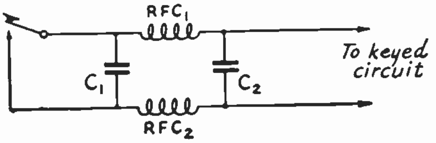

Fig. 8-8 — A filter for eliminating the r.f. click at the key. First try C_1, then add the two r.f. chokes, and then C_2. This filter does not eliminate on-the-air clicks, but it is necessary if you are trying to check keying in your own receiver. It should be mounted right at the key.

C_1, C_2 — 0.01 to 0.001 μfd., not critical.
RFC_1, RFC_2 — 1- to 2.5-mh. r.f. choke.

So far you haven't done a thing for your signal on the air and you still don't know what it sounds like, but you may have cleaned up some clicks in the b.c. set. Now disconnect the antenna from your receiver and short the antenna terminals with a short piece of wire. Tune in your own signal and reduce the r.f. gain to the point where your receiver doesn't overload. Detune any antenna trimmer the receiver may have. If you can't avoid overload within the r.f. gain-control range, pull out the r.f. amplifier tube and try again. If you still can't avoid overload, listen to the second harmonic as a last resort. Since an overloaded receiver can generate clicks, it is easy to realize the importance of eliminating overload during any tests or observations.

Describing the volume level at which you should set your receiver for these "shack" tests is a little difficult. The r.f. filter should be effective with the receiver running wide open and with an antenna connected. When you turn on the transmitter and take the other steps mentioned to reduce the signal in the receiver, run the audio up and the r.f. down to the point where you can just hear a little "rushing" sound with the b.f.o. off and the receiver tuned to the signal. This is with the crystal filter in. At this level, a properly-adjusted keying circuit will show no clicks off the rushing-sound range. With the b.f.o. on and the same gain setting, there should be no clicks outside the beat-note range. When observing clicks, make the slow-dash and fast-dot tests outlined previously.

Now you know how your signal sounds on the air, with one exception. If keying your transmitter makes the house lights blink or the dial light in your receiver flicker, you may not be able to tell too accurately about any chirp on your signal. However, if you are satisfied with the absence of chirp when tuning *either side of zero beat,* it is safe to assume that your receiver isn't chirping with the light flicker and the observed signal is a true representation. No chirp either side of zero beat is fine — some chirp can be either in your transmitter or your receiver, when the lights flicker. But don't try to make these tests without first getting rid of the r.f. click at the key — you will never be able to give yourself a clean bill of health, because clicks can mask a chirp.

In some instances, particularly if the transmitter power is several hundred watts or more, you may find that a small click still persists on all frequencies. If such a click is observed, pull out the last i.f. amplifier tube in your receiver and listen again. If the click is still there, it indicates rectification in the audio system of your receiver, the same type of BCI we cuss out cheap midget receivers for. You can cure it with the usual resistor-condenser filter used for curing such BCI cases, or you can leave it in and make mental compensation for it. Any click you hear on your signal should reduce to this minimum click immediately off the signal.

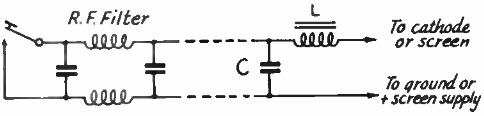

Fig. 8-9 — A key-click filter for cathode, negative-lead or screen keying. It can be located anywhere in the keying line. The values of L and C will vary widely with different currents and voltages, and must be found by cut-and-try. For screen keying, the resistor R_2 (Fig. 8-2) should connect to the junction of L and C.

C — 0.05 to 2.0 μfd.
L — 0.5 to 30 henrys.

Another unavoidable click can be encountered by r.f. pick-up on the lead from a receiver i.f. amplifier to a Q5-er. Here again the click will be present at any setting of the receiver tuning control. The solution here is to make your checks with the Q5-er disconnected and the lead removed from the receiver.

Key clicks are caused by the key turning your transmitter on and off too fast — and sometimes by parasitic oscillations in an amplifier — and all a key-click filter does is to slow down the turning-on and turning-off processes. Parasitic clicks occur at points 25 to 100 kc. either side of the signal, and are caused by low-frequency parasitic oscillations that are triggered by the keying. The cure consists of eliminating the oscillation, not adding key-click filters.

Plate, screen or cathode keying requires a key-click filter of the type shown in Fig. 8-9. Adjustment of such a filter is a simple matter. If the signal has too heavy a click or thump on "make," L should have more inductance. If the click is too heavy on "break," C should have more capacity. The "break" characteristic is also influenced by the value of L, so start with a value of C that reduces the clicks noticeably on "break," adjust the value of L for best "make" characteristic, and then clean up the "break" by further modification of C. Since you may have only a few stray inductances around the shack, you may not find just the value you want for L. In this case, use a value that gives too soft a "make" and then shunt the inductance with resistance to reduce its effect. Transformer windings will often serve as well as standard chokes in this application, so try everything around the shack until you find what you need. For a given voltage, high-current circuits will require more C and less L than will low-current ones.

In the screen-grid keying circuit, the value of R_2 will also affect the "break" characteristic. If R_2 is too large the "break" will tail off too gradually, if it is too small it may introduce a click on "break." In general it is best to start with a value as suggested in Fig. 8-2 and adjust C for the proper "break" characteristic.

Adjustment of control-grid or keyer-tube keying characteristics is simple, since the important components are C_1, R_1 and R_2 (Figs. 8-3 and 8-5). For a given value of C_1, increasing the value of R_2 will soften the "make" characteristic, and increasing the value of R_1 will soften the "break." The value of R_1 will be many times the value of R_2. With grid-block keying, the value of R_2 is determined already if the tube runs grid current, because this will be the normal grid leak, and so the value of C_1 must be adjusted for proper "make" characteristic and then the "break" made satisfactory by adjustment of R_1. Tubes running heavy grid current are not too suitable for grid-block keying because the value of R_1 generally ends up comparatively low and the negative supply must furnish too much current when the key is down.

If you are keying in a low-level stage, don't overlook the clipping action of subsequent stages that are fixed-biased beyond cut-off. It can reintroduce clicks.[2] And if you key your oscillator, don't be too disappointed in the chirp that shows up when you have clickless keying. Amplifier keying is the answer.

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[2] For a more complete discussion of this effect, see Carter, "Reducing Key Clicks," *QST*, March, 1949.

A Vacuum-Tube Keyer

A tube-keyer unit is shown in Figs. 8-10 and 8-11. 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 connected 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.

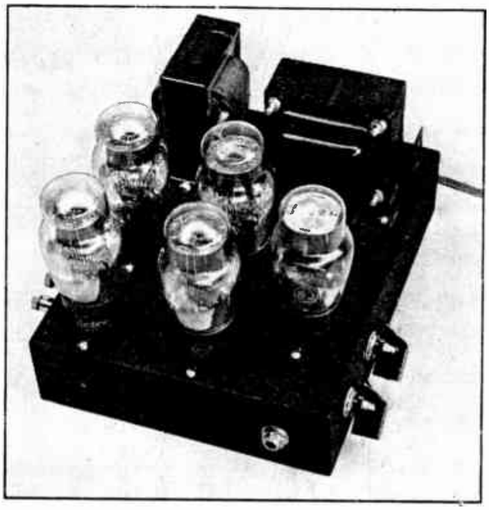


Fig. 8-10 — A vacuum-tube keyer, built up on a 7 × 9 × 2-inch chassis with space for four or less keyer tubes and the power-supply rectifier. The resistors and condensers that produce the lag are underneath, controlled by the knobs at the right. The jack is for the key, while terminals at the left are for the keyed circuit.

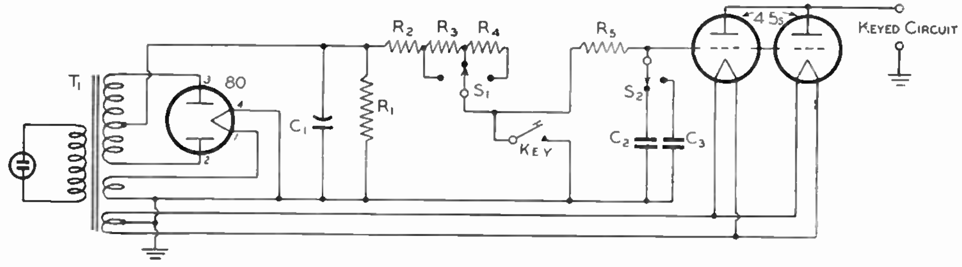


Fig. 8-11 — Wiring diagram of a practical vacuum-tube keyer similar to the one in Fig. 8-10.

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 — 350-0-350 volts, 5 volts and 2.5 volts (Stancor P6003).

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. 8-12 and 8-14. 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. 8-13. 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

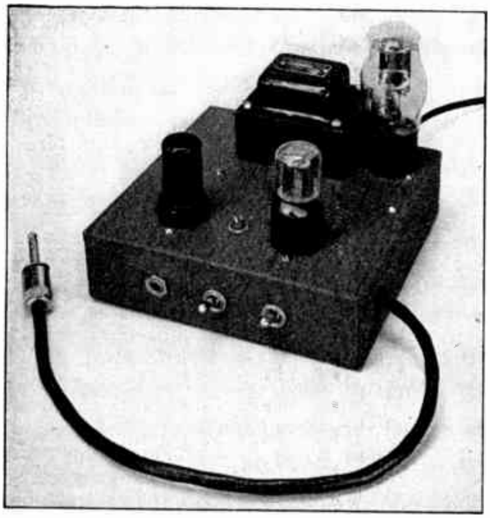


Fig. 8-12 — A top view of the "Monitone." The shaft of the screwdriver-adjusted potentiometer controlling tone and volume is located between the 6J5 and 6SL7 tubes. The right-hand switch controls the a.c., and the center switch cuts the tone oscillator in and out.

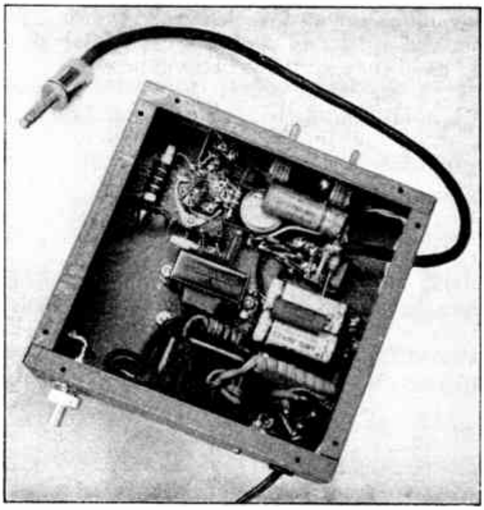


Fig. 8-14 — Bottom view of the "Monitone," showing the small neon lamp supported on its own leads and located directly above the center of the 6J5 socket. The r.f. pick-up line terminates on the tie-point that carries the 1N34.

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. 8-12 and 8-14. 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-

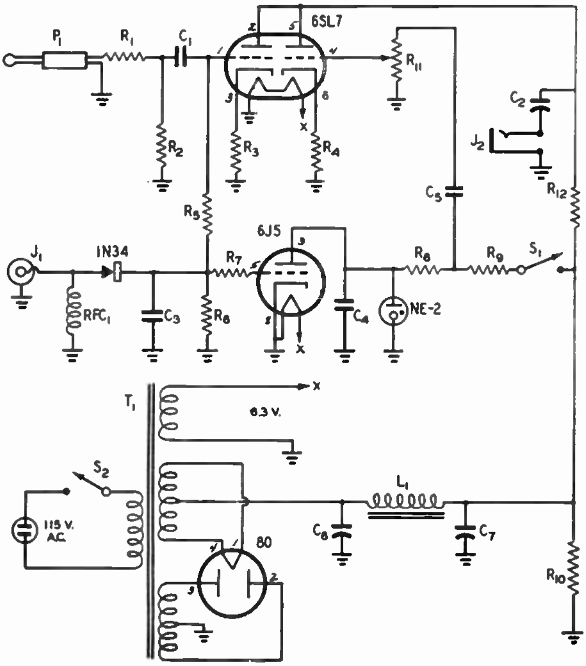


Fig. 8-13 — Circuit diagram of the "Monitone."

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 — 10-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).

up wire, either directly from the Monitone or extending beyond the shielding of the coaxial 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.

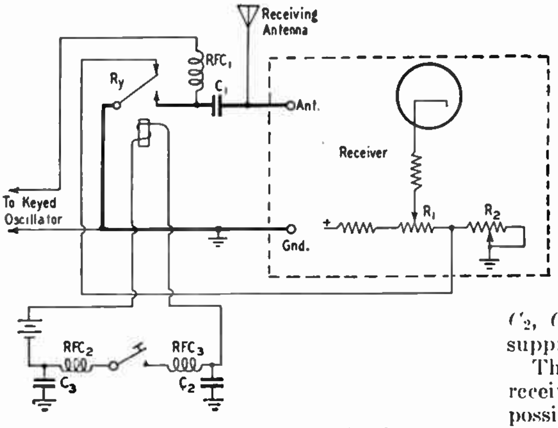


Fig. 8-15 — Wiring diagram for smooth break-in operation. The leads shown as heavy lines should be kept as short as possible, for minimum pick-up of the transmitter signal.

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. 8-15 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. 8-16.

● 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 if the rise and decay times are made very short, but this introduces key clicks that cannot be

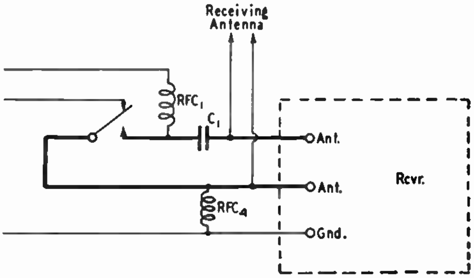


Fig. 8-16 — Necessary circuit revision of Fig. 8-15 if a two-wire lead from the receiving antenna is used. RFC_4 is a 2.5-mh. r.f. choke — other values are the same as in Fig. 8-15.

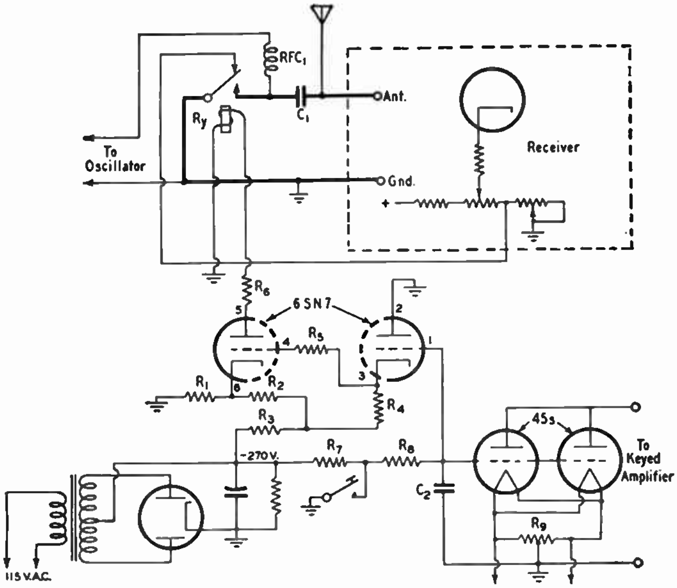


Fig. 8-17 — A de luxe break-in system that holds the oscillator circuit closed (and the receiver input shorted) during a string of fast dots but opens between letters or words.

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).

avoided. The system shown in Fig. 8-17 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. 8-17, the circuit is a combination of the break-in system of Fig. 8-15 and the tube keyer of Fig. 8-11, 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. By using blocked-grid keying of the amplifier stage, the keyer tubes can be eliminated.

● ELECTRONIC KEYS

Electronic keys, as contrasted with mechanical automatic keys, use vacuum tubes or relays (or both) to form automatic dashes as well as automatic dots. As first devised by amateurs in 1940, a dash could be "clipped short" if the dash lever were lifted too soon. More recent designs have resulted in "self-completing dashes" that eliminate this possibility and permit the operator, with a reasonable amount of practice, to generate near-perfect code. Full descriptions of electronic keys that produce self-completing dashes can be found in the following QST articles:

Bartlett, "Further Advances in Electronic-Keyer Design," October, 1948; correction, page 10, January, 1949.

Turrin, "Debugging the Electronic Bug," Jan., 1950.

Montgomery, "'Corkey' — A Tubeless Automatic Key," November, 1950.

A simple unit that can be attached to a mechanical automatic key to give automatic dashes (not of the self-completing type, however) can be found described in the following *QST* article:

Gotisar, "The Dash Master," Aug., 1948.

CHAPTER 9

Speech Amplifiers And Modulators

The audio amplifiers used in radiotelephone transmitters operate on the principles outlined earlier in this book in the chapter on vacuum tubes. The design requirements are determined principally by the type of modulation system to be used and by the type of microphone to be employed. It is necessary to have a clear understanding of modulation principles before the problem of laying out a speech system can be approached successfully. Those principles are discussed under appropriate chapter headings.

The present chapter deals with the design of audio amplifier systems for communication purposes. In voice communication the primary objective is to obtain the most *effective* transmission; i.e., to make the message be understood at the receiving point in spite of adverse conditions created by noise and interference. The methods used to accomplish this do not necessarily coincide with the methods used for other purposes, such as the reproduction of music or other program material. In other words, "naturalness" in reproduction is distinctly secondary to intelligibility.

The fact that satisfactory intelligibility can be maintained in a relatively narrow band of frequencies is particularly fortunate, because the width of the channel occupied by a 'phone transmitter is directly proportional to the width of the audio-frequency band. If the channel width is reduced, more stations can occupy a given band of frequencies without mutual interference. Also, it is possible to use more selectivity in the receiver and thus effect a further improvement.

In speech transmission, amplitude distortion of the voice wave has very little effect on intelligibility. Its importance in communication lies almost wholly in the fact that the audio-frequency harmonics caused by such distortion may lie outside the channel needed for intelligible speech, and thus will create unnecessary interference to other stations.

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.

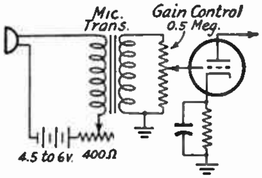
● MICROPHONES

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.

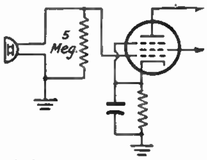
The **frequency response** or **fidelity** of a microphone is its relative ability to convert sounds of different frequencies into alternating current. With fixed sound intensity at the microphone, the electrical output may vary considerably as the sound frequency is varied. For understandable speech 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

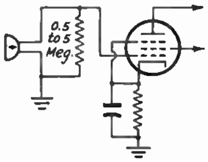
The **carbon microphone** consists of a metal diaphragm placed against an insulating cup containing loosely-packed carbon granules (**microphone button**). Current from a battery flows through the granules, the diaphragm be-



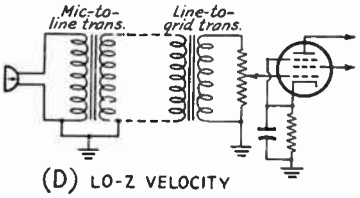
(A) S.B CARBON



(B) CRYSTAL



(C) HI-Z VELOCITY



(D) LO-Z VELOCITY

Fig. 9-1 — Speech input circuits used with various types of microphones.

ing one connection and the metal backplate the other. Fig. 9-1A 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. 9-1B.

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. 9-1C). 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. 9-1D.

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 suitable modulator for the transmitter. This selection must be based on the power required to modulate the transmitter, and this power in turn depends on the type of modulation system selected, as described in other chapters. 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 the last chapter. 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.

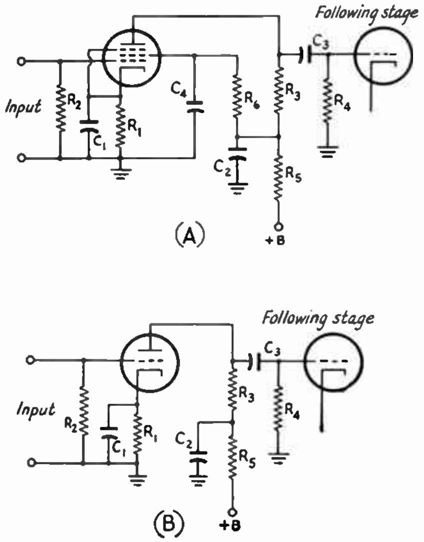


Fig. 9-2 — Resistance-coupled voltage-amplifier circuits. A, pentode; B, triode. Designations are as follows:

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.

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. 9-2 and design data in Table 9-1.

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 at the end of this book.

Representative circuits for coupling single-ended to push-pull stages are shown in Fig. 9-3. The circuit at A combines resistance and transformer coupling, and may be used for exciting the grids of a Class A or AB₁ 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.

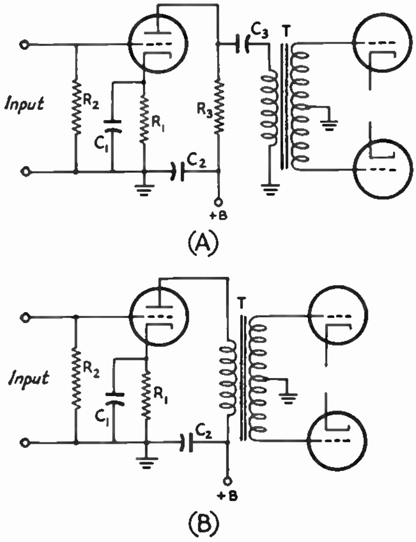


Fig. 9-3 — Transformer-coupled amplifier circuits for driving a push-pull amplifier. A is for resistance-transformer coupling; B for transformer coupling. Designations correspond to those in Fig. 9-2. In A, values can be taken from Table 9-I. In B, the cathode resistor is calculated from the rated plate current and grid bias as given in the tube tables for the particular type of tube used,

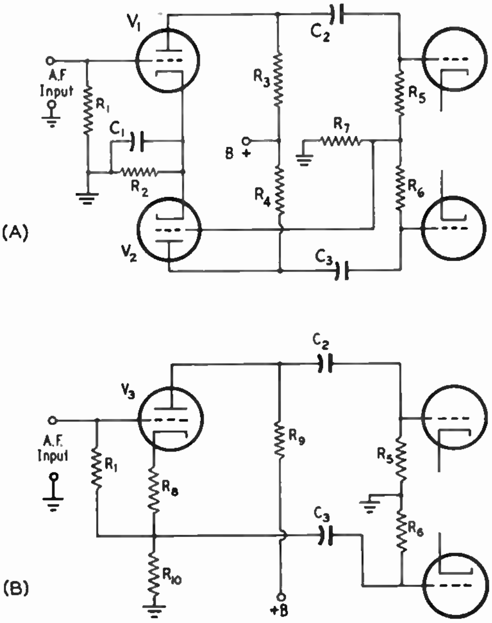


Fig. 9-4 — Self-balancing phase-inverter circuits. V_1 and V_2 may be a double triode such as the 6SN7GT or 6SL7GT. V_3 may be any of the triodes listed in Table 9-I, or one section of a double triode.

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 AB₂ or Class B stage.

Phase Inversion

Push-pull output may be secured with resistance coupling by using "phase-inverter" circuits as shown in Fig. 9-4.

The circuit shown in Fig. 9-4A 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 9-1—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]
6SJ7,12SJ7	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, 12J7-GT	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, 12AU6, 12SH7	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, 6AQ7, 6AT6, 6Q7, 6SL7GT, 6SZ7, 6T8, 12AT6, 12Q7-GT, 12SL7-GT (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
6AV6, 12AV6, 12AX7 (one triode)	0.1	0.1	—	1300	—	4.6	0.027	43	45
		0.22	—	1500	—	4.0	0.013	57	52
		0.47	—	1700	—	3.6	0.006	66	57
	0.22	0.22	—	2200	—	3.0	0.013	54	59
		0.47	—	2800	—	2.3	0.006	69	65
		1.0	—	3100	—	2.1	0.003	79	68
	0.47	0.47	—	4300	—	1.6	0.006	62	69
		1.0	—	5200	—	1.3	0.003	77	73
		2.2	—	5900	—	1.1	0.002	92	75
6SC7, 12SC7 [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, 12J5-GT, 12SN7-GT (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, 12AU7 (one triode)	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. 9-4B 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 at the end of this book the grid 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 or output stage of the speech-amplifier. Receiver-type power tubes can be used (beam tubes such as the 6L6 may be needed in some cases) as determined from the receiving-tube tables. If the speech amplifier is to drive a Class B modulator, use a Class A or AB_1 amplifier, in preference to Class AB_2, if it will give enough power output.

4) If the speech-amplifier output 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 speech-amplifier output stage operates Class A or AB_1, it may be driven by a voltage amplifier. If the output 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. 9-4A) 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 described earlier in this chapter.

7) Divide the voltage required to drive the output 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.

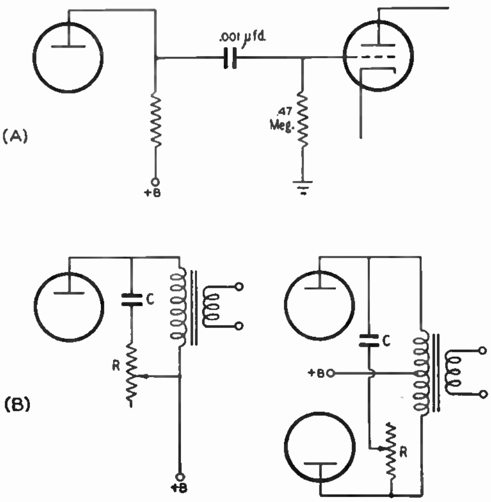


Fig. 9-5 — A, use of a small coupling condenser to reduce low-frequency response; B, tone-control circuits for reducing high-frequency response. Values for *C* and *R* are discussed in the text; 0.01 µfd. and 25,000 ohms are typical.

One simple way to reduce low-frequency response is to use small values of coupling capacitance between resistance-coupled stages, as shown in Fig. 9-5A. 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. 9-5B. 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. 9-6. 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

In the chapter on amplitude modulation it is 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

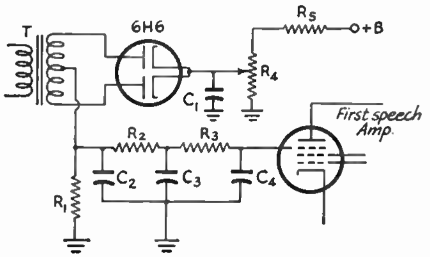


Fig. 9-6 — Speech-amplifier output-limiting circuit. C_1, C_2, C_3, C_4 — 0.1-µfd.; R_1, R_2, R_3 — 0.22 megohm; R_4 — 25,000-ohm pot.; R_5 — 0.1 megohm; T — see text.

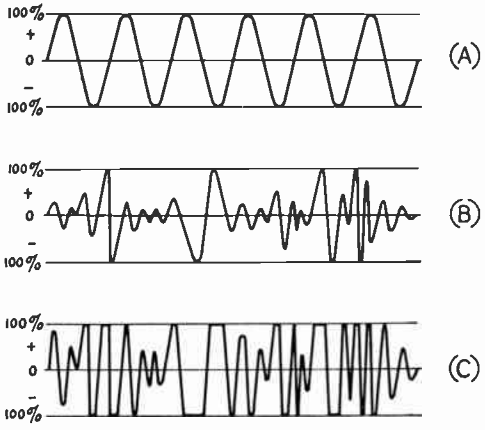


Fig. 9-7 — The normal speech wave (B) has high peaks but low average energy content. When the peaks are clipped the signal may be increased to a considerably-higher power level without causing overmodulation (C).

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. 9-7. 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 suffers; 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.

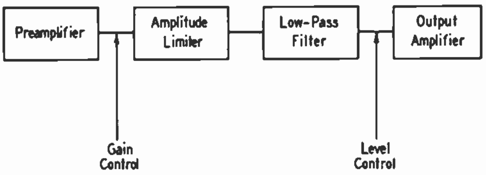


Fig. 9-8 — Block diagram of speech-clipping and filtering amplifier.

There is a loss in naturalness with "deep" clipping, even though the voice is highly intelligible. With moderate clipping levels (6 to 12 db.) there is almost no perceptible change in "quality" but the voice power is four or sixteen 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. 9-8. 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 or clipping 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" or modulation 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 in this chapter.

A Clipper-Filter Speech Amplifier-Driver

The speech amplifier shown in Figs. 9-9, to 9-11, inclusive, uses push-pull triodes to obtain a power output of 13 watts with negligible distortion — sufficient to drive most of the commonly-used Class-B modulator tubes. It includes a clipper-filter for increasing the effectiveness of modulation and for confining the channel width to frequencies needed for intelligible speech. The over-all gain is ample for use with communications-type crystal microphones when using clipping of the order of 12–15 db. Miniature tubes are used in the voltage-amplifier stages. The output tubes are 6B4Gs, operated Class AB1 with fixed bias. Two power supplies are included, one for the voltage amplifier stages and the other for the output tube plates.

As shown in Fig. 9-10, the first two stages are voltage amplifiers of ordinary design, using a 6AU6 pentode in the first stage and a 6C4 triode in the second. The output of the second stage can be switched either to the 12AU7 double-triode clipper or to the 6C4 voltage amplifier that drives the 6B4G grids. In the latter case the amplifier operation is conventional. The clipper, when operative, provides additional voltage gain as well as clipping. Its output goes through a simple low-pass filter ($L_1C_{11}C_{12}$) so that harmonics generated by clipping will be attenuated before the signal reaches the grid of the second 6C4. The frequency response of the amplifier with the filter in circuit, but with the signal below the clipping level, drops at the rate of roughly 6 db. per octave below 500 cycles; above 4000 cycles the response is down 25 db. compared with the medium audio range.

A two-section filter is used in the plate supply for the voltage-amplifier stages. The hum level must be kept low because of the high gain required when using clipping. A single-section filter is sufficient for the output stage. Bias for the 6B4G grids is obtained from the low-voltage supply by means of R_{16}, by-passed by C_{14}.

Two gain controls are included, one (R_6) for setting the level into the clipper circuit and thus determining the amount of clipping, and the second (R_{13}) for setting the output level after clipping. With the clipper in use, proper setting of R_{13} will keep the modulation level high but will prevent overmodulation.

Construction

As shown in Fig. 9-9, the voltage amplifiers occupy the left front section of the chassis. The 6AU6 first amplifier is at the left, followed in order to the right by the first 6C4, the 12AU7, and the second 6C4. The 6B4Gs and their output transformer are at the right front. The cylindrical unit just behind the second 6C4 is the interstage audio transformer, T_1.

Power supply components are grouped along the rear edge of the chassis, with the low-voltage supply at the left. The power transformers should be kept well separated from the voltage amplifiers, particularly the first two stages, in order to minimize hum difficulties.

On the front panel, the microphone input connector is at the lower left. Next to it is the clipping control, then the clipper in-out switch, and then the modulation control. The two toggle switches at the right are S_2 and S_3. The a.c. input socket is by-passed by C_{15} and C_{16}, to reduce the possibility that r.f. picked up on the line cord will get into the low-level speech stages.

The wiring underneath the chassis is relatively simple, as shown by Fig. 9-11. The microphone input circuit, including RFC_1 and C_1, is enclosed in a National jack shield, and the lead from RFC_1 to the 6AU6 grid also is shielded.

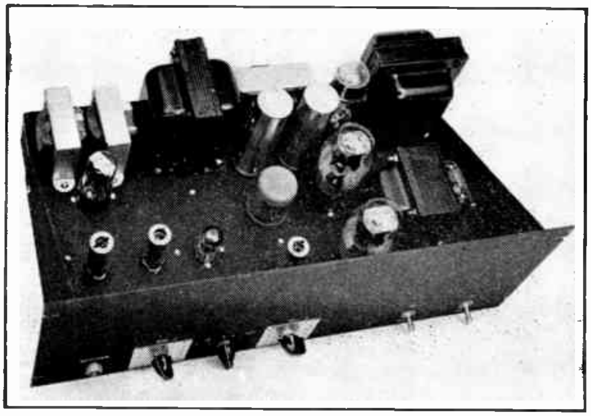
Adjusting the Clipper-Filter Amplifier

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. 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.

◆

Fig. 9-9 — This speech-amplifier and driver has ample gain for a crystal microphone and is complete with power supply. The measured undistorted output is 13 watts. It incorporates a clipper-filter system for increasing modulation effectiveness and decreasing channel width.

◆



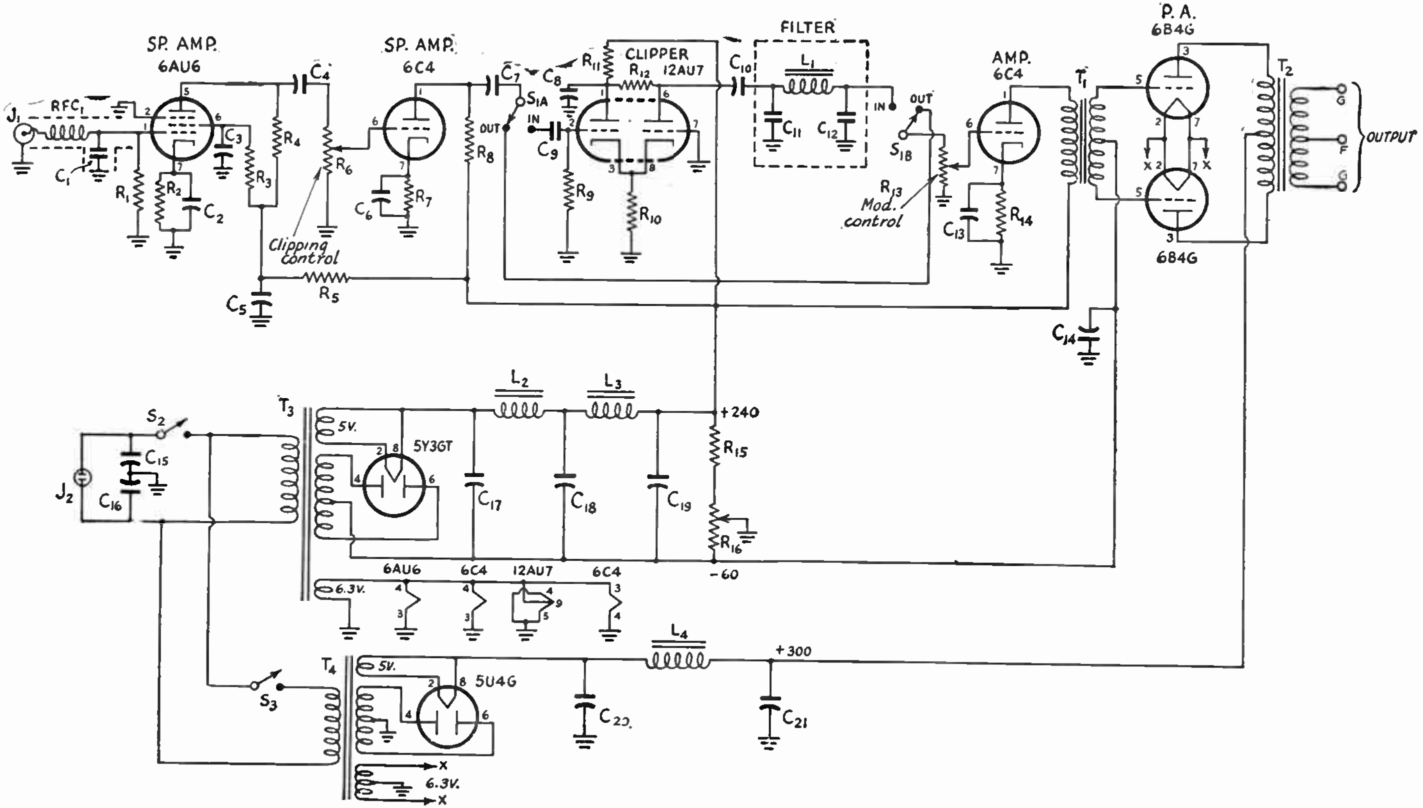


Fig. 9-10 — Circuit diagram of the clipper-filter speech amplifier.

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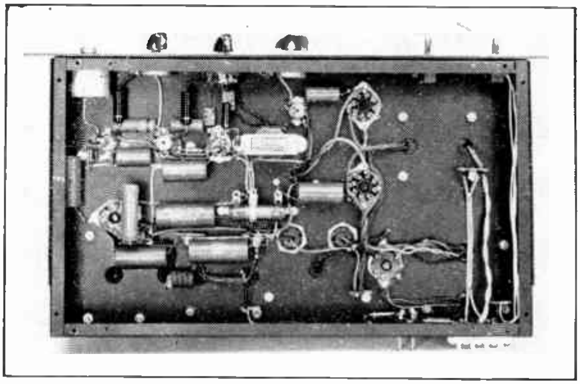

Fig. 9-11 — Below-chassis view of the clipper-filter speech amplifier. The relatively small number of components below the chassis makes wiring simple.

◆

The setting of R_{13} 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 (3000 ohms) at the plates of the 6B4Gs. First set R_{13} 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_1 to the "normal" or "out" position; the waveshape should return to normal. If it does not, return S_1 to the "in" position and reduce the setting of R_{13} until it does. Then reduce the amplifier gain by means of R_6 until the signal is just below the clipping level. At this point the signal should be a sine wave. Increase R_{13}, without touching R_6, until the wave starts to become distorted, and then back off R_{13} until distortion disappears.

Next, change the input-signal frequency to 2500 cycles, without changing the signal level. Slowly increase R_6 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_{13} until the distortion disappears, even when R_6 is set at maximum and the maximum available signal from the audio oscillator is applied to the amplifier. The position of R_{13} should be noted at this point and the observed setting should never be exceeded.

To find the *operating* setting of R_{13}, 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_{13} (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_{13} 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_6, so that the transmitter is modulated 100 per cent. Observe the pattern closely at different settings of R_6 to see if it is possible to overmodulate. If overmodulation does not occur at any setting of R_6, the transmitter is ready for operation and R_{13} may be locked in position; it need never be touched subsequently. If some overmodulation does occur, R_{13} 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.

C_1 — 100-$\mu\mu$fd. mica.
C_2, C_6, C_{13} — 20-μfd. 25-volt electrolytic.
C_3 — 0.1-μfd. 400-volt paper.
C_4, C_7, C_{15}, C_{16} — 0.01-μfd. 400-volt paper.
C_5, C_8 — 8-μfd. 450-volt electrolytic.
C_9, C_{11} — 470-$\mu\mu$fd. mica.
C_{10} — 0.002-μfd. mica or paper.
C_{12} — 330-$\mu\mu$fd. mica.
C_{14} — 30-μfd. 150-volt electrolytic.
C_{17}, C_{18}, C_{19} — 16-μfd. 450-volt electrolytic.
C_{20}, C_{21} — 8-μfd. 450-volt electrolytic (can type).
R_1 — 2.2 megohms, ½ watt.
R_2, R_{14} — 2200 ohms, ½ watt.
R_3 — 1 megohm, ½ watt.
R_4, R_9 — 0.47 megohm, ½ watt.
R_5 — 47,000 ohms, ½ watt.
R_6 — 2-megohm volume control.
R_7 — 3900 ohms, ½ watt.
R_8 — 0.1 megohm, ½ watt.
R_{10} — 1500 ohms, 1 watt.
R_{11} — 47,000 ohms, 1 watt.
R_{12} — 56,000 ohms, ½ watt.
R_{13} — 0.5-megohm volume control.
R_{15} — 10,000 ohms, 20 watts.
R_{16} — 2000-ohm 25-watt adjustable.
L_1 — 20 henrys, 900 ohms (Stancor C-1515).
L_2, L_3 — 15 henrys, 75 ma. (Stancor C-1002).
L_4 — 10.5 henrys, 110 ma. (Stancor C-1001).
J_1 — Microphone cable receptacle (Amphenol PC1M).
J_2 — Chassis-mounting 115-volt plug.
S_1 — D.p.d.t. rotary switch (Mallory 3122-J).
S_2, S_3 — S.p.s.t. toggle.
T_1 — Audio transformer, single plate to p.p. grids, ratio 2:1 (Thordarson T20A17).
T_2 — Driver transformer, variable ratio, p.p. driver to Class-B grids, pri. rating 120 ma. per side (Stancor A-4763).
T_3 — Power transformer: 700 v. c. t., 90 ma.; 5 v., 2 amp.; 6.3 v. 3.5 amp. (Stancor P-4079).
T_4 — Power transformer: 700 v. c. t., 110 ma.; 5 v., 3 amp.; 6.3 v. 4.5 amp. (Stancor P-4080).
RFC_1 — 2.5 mh. r.f. choke.

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. 9-12 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, 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.

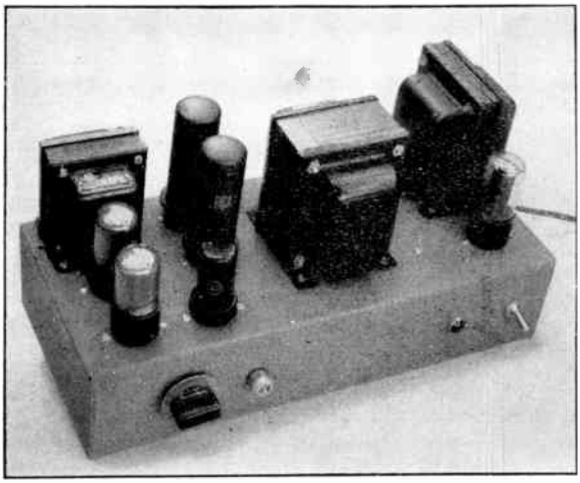

Fig. 9-12 — A 40-watt modulator of inexpensive construction. The second tube from the left, in the foreground, is the 6SJ7 first amplifier. The microphone connector is immediately below it on the chassis wall. Along the left edge, from the front, are the first and second 6SN7GTs and the driver transformer for the 6L6s. The output transformer is to the right of the 6L6s. The power transformer and rectifier are at the far right.

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 should be adjusted so the voltage drop across it is 22.5 volts when the speech-amplifier stages are taking normal current.

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 the power-supply chapter.

20-Watt Modulator

Fig. 9-15 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

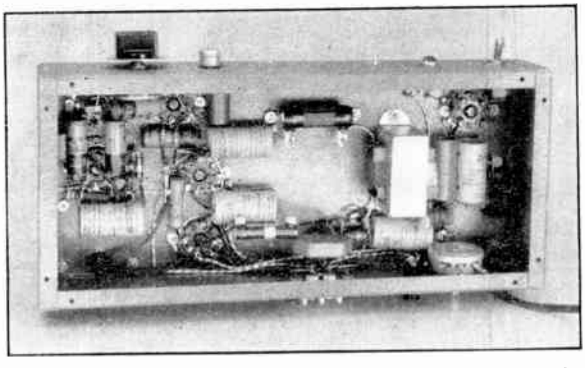

Fig. 9-13 — Underneath the chassis of the 40-watt modulator. The power-supply choke is mounted below chassis at the right. The bias-setting resistor, R_{19}, is on the rear chassis wall, at the lower right in this photograph. Other components are grouped near the tube socket with which they are associated.

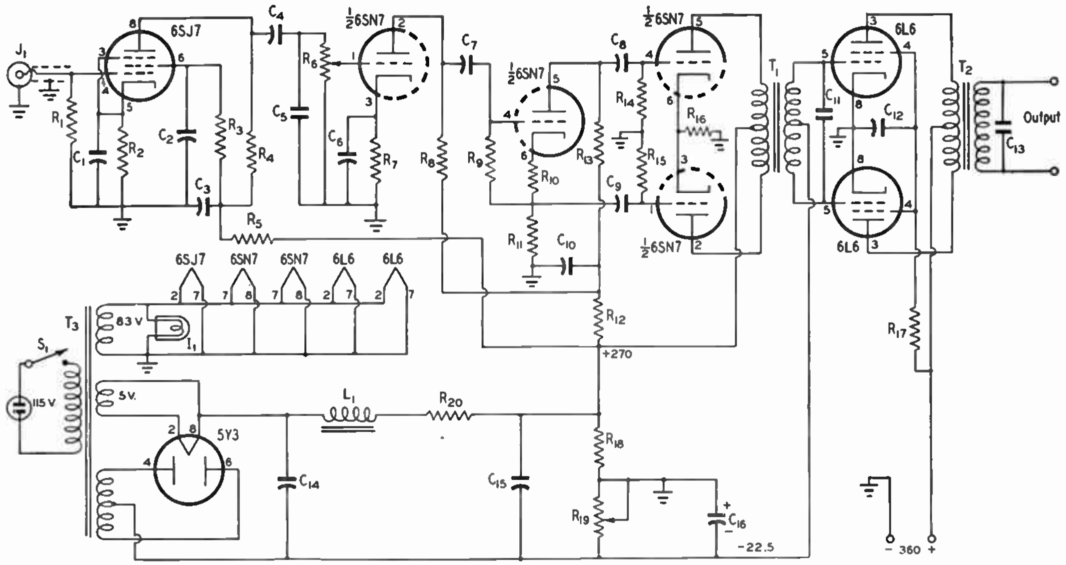

Fig. 9-14 — Circuit diagram of the 40-watt modulator.

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 AB_2 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 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.

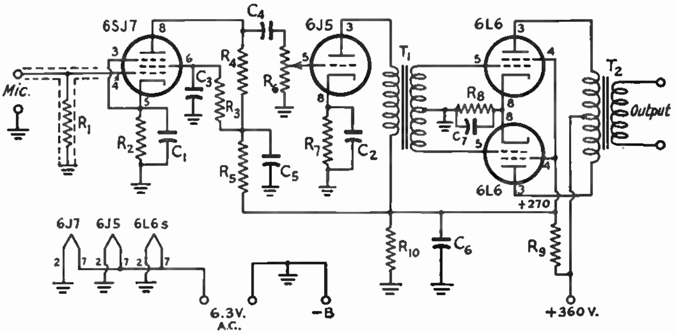

Fig. 9-15 — Circuit diagram of a low-cost modulator capable of power outputs up to 20 watts.

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. 9-16 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. 9-17, 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

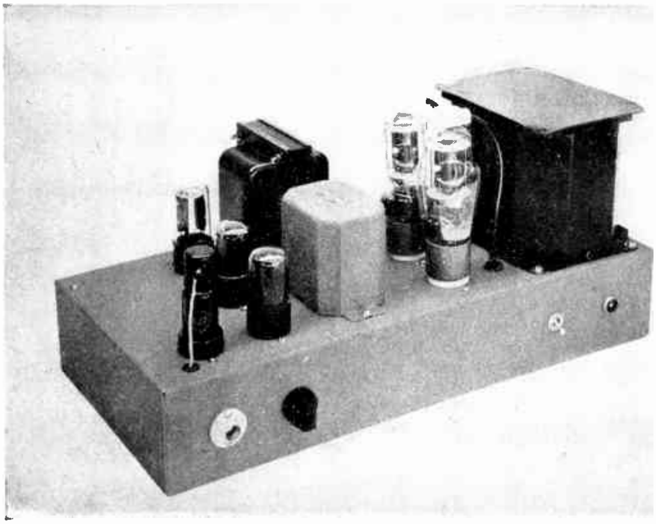

Fig. 9-16 — A speech amplifier and 807 modulator for plate modulation of transmitters up to 200 watts input. The microphone jack and the gain control are at the left end of the chassis. The audio components and tubes occupy the front section, and the power supply for the driver tubes is laid out along 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.

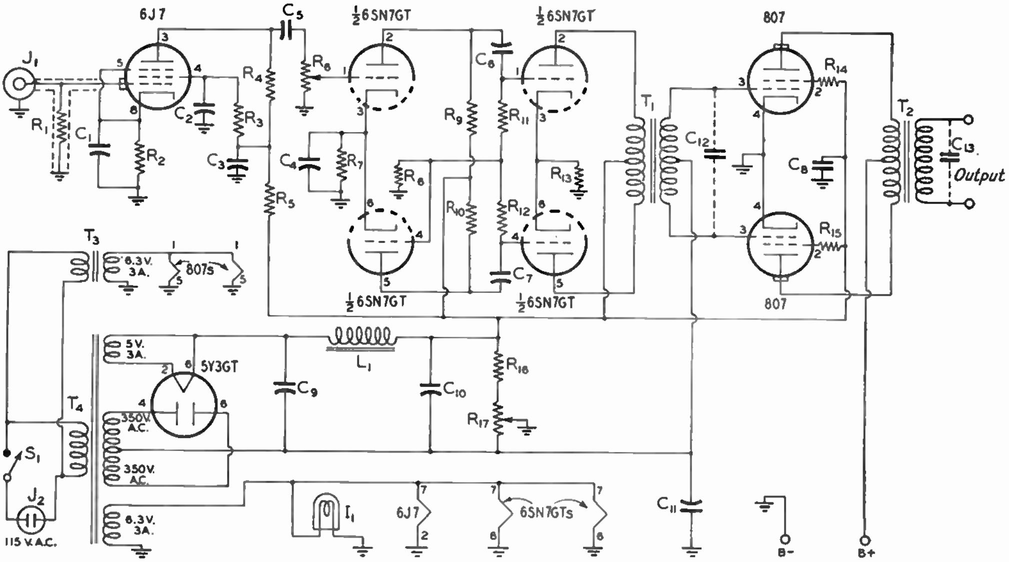

Fig. 9-17 — Circuit diagram of the push-pull 807 speech amplifier-modulator.

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.

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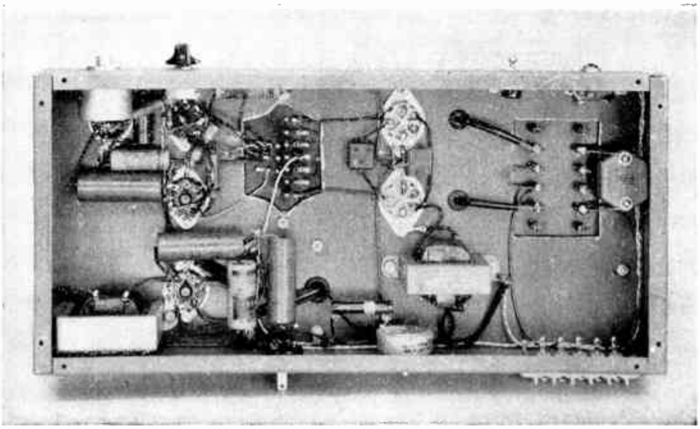
Fig. 9-18 — Below-chassis view of the 807 modulator. The shielded microphone jack is in the upper left-hand corner. The filter choke is mounted in the lower left-hand corner and the 807 filament transformer is to the rear and slightly to the right of the 807 tube sockets. The condenser for attenuating 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.



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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. 9-19 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 chapter containing the tube tables. 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.

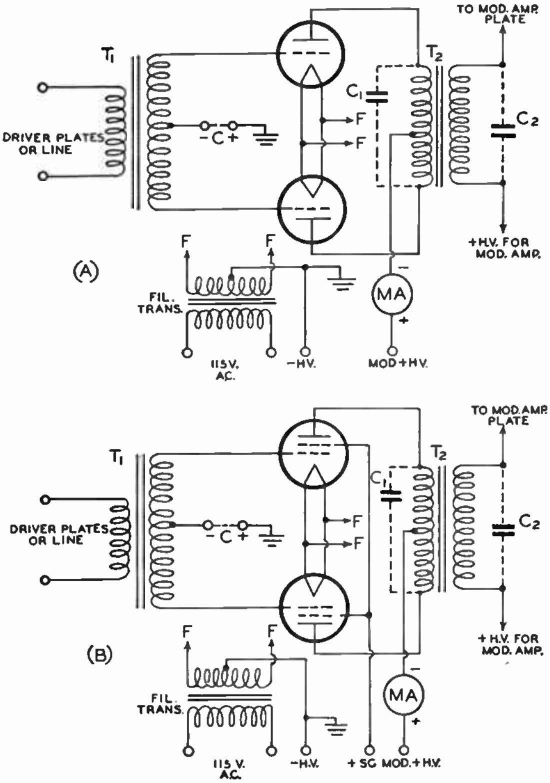

Fig. 9-19 — Class B modulator circuit diagrams. Tubes and circuit considerations are discussed in the text.

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

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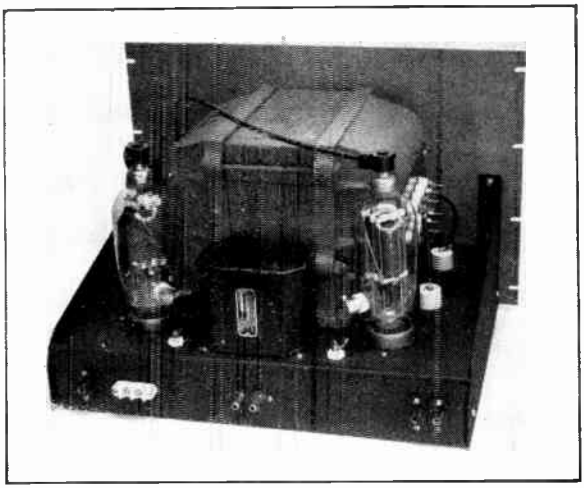

Fig. 9-20 — A typical chassis layout for a Class B modulator. Beyond adequate insulation for the voltages used, and sufficient ventilation for the modulator tubes, no particular constructional precautions are necessary. If the size of the components makes it necessary to use more than one chassis, the 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.

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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. 9-19 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 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 the power supply chapter.

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. 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. The audio impedance between screen and cathode also must be low.

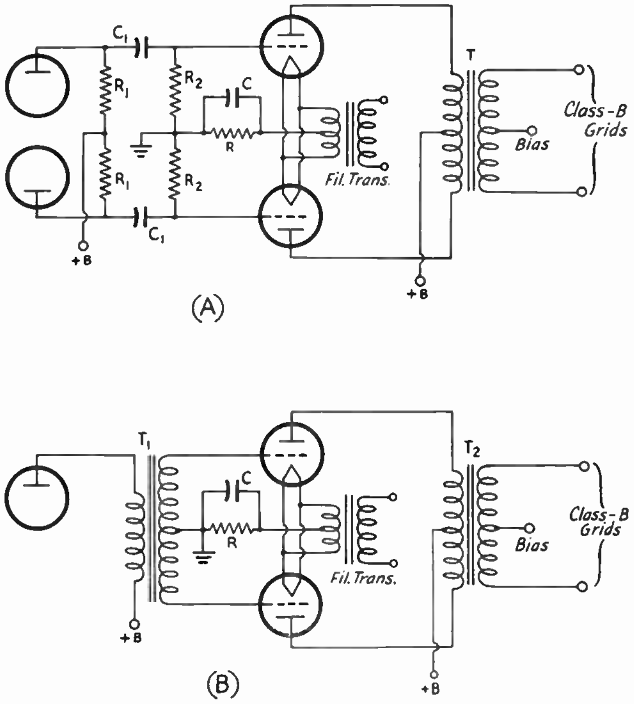


Fig. 9-21 — Triode driver circuits for Class B modulators. A, resistance coupling to grids; B, transformer coupling. R_1 in A is the plate resistor for the preceding stage, value determined by the type of tube and operating conditions as given in Table 9-I. C_1 and R_2 are the coupling condenser and 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.

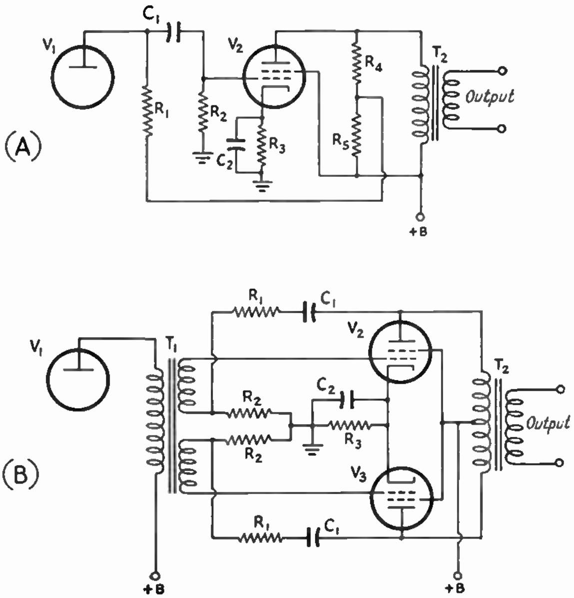


Fig. 9-22 — Negative feed-back circuits for drivers for Class B modulators. A — Single-ended beam-tetrode driver. If V_1 and V_2 are a 6J5 and 6V6, respectively, the following values are suggested: R_1, 47,000 ohms; R_2, 0.47 megohm; R_3, 250 ohms; R_4, R_5, 22,000 ohms; C_1, 0.01 μfd.; C_2, 50 μfd.

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 harmonics 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. 9-21, 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 6B4G 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

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. 9-21 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 6B4Gs 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 2A3 or 6B4G.

Suitable circuits for single-ended and push-pull tetrodes are shown in Fig. 9-22. Fig. 9-22A 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. 9-22.

The push-pull circuit in Fig. 9-22B 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

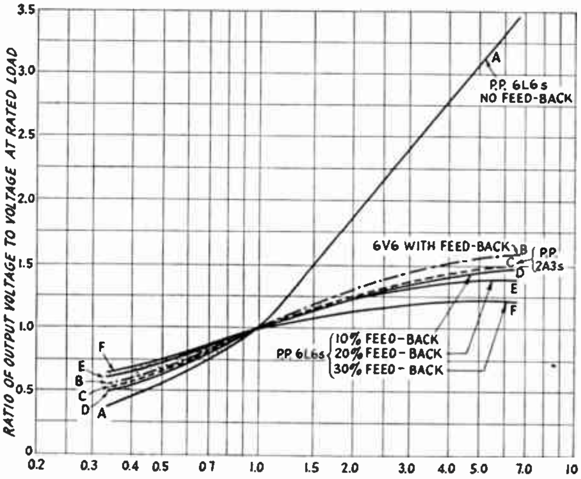


Fig. 9-23 — Output voltage regulation of two types of beam-tetrode drivers with negative feed-back. For comparison, the regulation with a pair of 2A3s (no feed-back) also is shown.

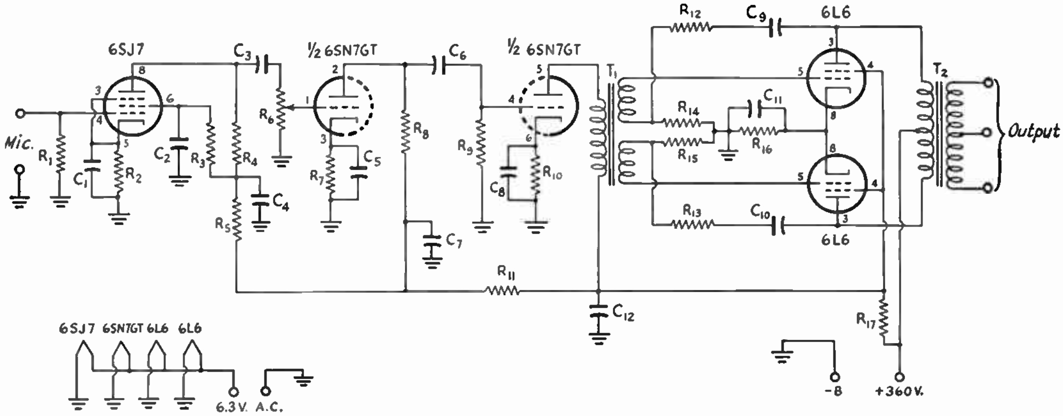


Fig. 9-24 — Circuit diagram of speech amplifier using 6L6s with negative feed-back, suitable for driving Class B modulators up to 500 watts output.

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.

AB₂. 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 AB₁ 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 6B4Gs or 2A3s without feed-back.

Instead of the voltage-divider arrangement shown in Fig. 9-22B 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. 9-23. 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. 9-24. In this amplifier the 6L6s are operated Class AB₁ 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. 9-12, 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. 9-15, 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 AB₁ 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. 9-15.

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. 9-25. The only equipment that is not likely to be already at hand is the audio oscillator (their construction is described in the chapter on measurements). 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. 9-25 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.

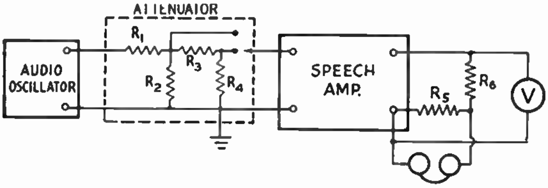


Fig. 9-25 — Simple test set-up for checking a speech amplifier. The audio-oscillator frequency range should be from about 100 to 5000 or more cycles. It is not necessary that it be continuously variable; a number of "spot" frequencies will be satisfactory. Suitable resistor values are: R_1 and R_3, 10,000 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. 9-26. 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. 9-27, 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. 9-27. 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 appreciable phase shifts. It is therefore advisable to test for distortion with an input signal that is as nearly as possible a sine wave. Also, it is best to use a frequency in the 500–1000 cycle range, since improper phase shift in the amplifier is usually least in this region. Phase shift in itself is not of much importance in an audio amplifier of ordinary design, because it does not change the character of speech so far as the ear is concerned, but if a complex signal is used for testing it is sometimes difficult to detect distortion in the oscilloscope pattern.

In amplifiers having negative feed-back, excessive phase shift may cause self-oscillation. This is because internal phase shift in one or more stages causes the signal fed back to aid, instead of oppose, the input signal. Oscillation usually occurs at some frequency above 10,000 cycles, and is associated with phase shift in a transformer, although occasionally it will occur at a very low frequency. It can be prevented by reducing the gain at the frequency at which it occurs. If the pass-band is deliberately restricted to the optimum voice range, as described earlier, the gain at both very high and very low frequencies will be so low that self-oscillation is very unlikely, even with large amounts of feed-back.

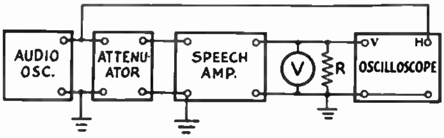


Fig. 9-26 — **Test set-up using the oscilloscope to check for distortion. These connections will result in the type of pattern shown in Fig. 9-27, the horizontal sweep being provided by the audio input signal. For waveform patterns, omit the connection between the audio oscillator and the horizontal amplifier in the 'scope, and use the horizontal linear sweep.**

Generally speaking, it is easier to detect small amounts of distortion with the type of pattern shown in Fig. 9-27 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.

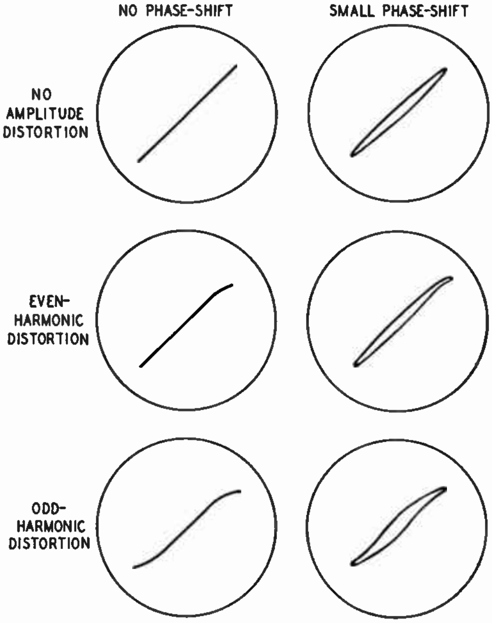


Fig. 9-27 — **Typical patterns obtained with the connections shown in Fig. 9-26. Depending on the number of stages in the amplifier, the pattern may slope upward to the right, as shown, or upward to the left. Also, depending on where the distortion originates, the curvature in the second row may appear either at the top or bottom of the line or ellipse.**

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.

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. 9-28. 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.

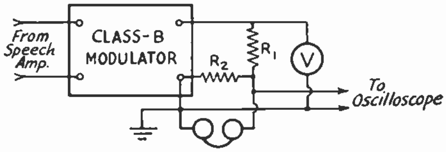


Fig. 9-28 — Set-up for checking a Class B modulator.

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.

CHAPTER 10

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.

There are several basic ways in which the carrier can be modulated. In the most common method, **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.

● SIDEBANDS

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 they are numerically equal to the carrier frequency plus the modulating frequency and the carrier frequency minus the modulating frequency. 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 maximum frequency in the modulating signal.

As explained in the chapter on audio equipment, it is unnecessary to transmit audio frequencies greater than about 3000 cycles for the intelligible reproduction of average speech. The maximum channel that it is necessary to use for voice communication, therefore, is twice 3000, or 6 kc. wide. If higher frequencies are transmitted the channel width is increased proportionately, thus creating unnecessary interference with stations operating on adjacent channels.

● THE MODULATED WAVE

In Fig. 10-1, 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

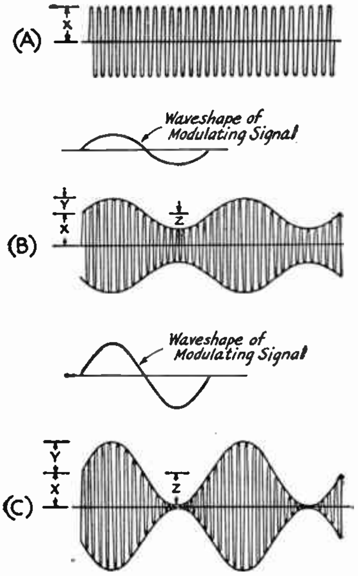


Fig. 10-1 — Graphical representation of (A) carrier unmodulated, (B) modulated 50%, (C) modulated 100%.

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 loudspeaker 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. 10-1C 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. 10-1, 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. 10-1C, 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. 10-1 and 10-2.

Power in Modulated Wave

The amplitude values shown in Fig. 10-1 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. 10-1C 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. 10-2 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.

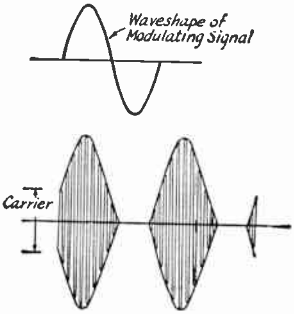


Fig. 10-2 — An overmodulated r.f. carrier wave.

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."

GENERAL 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. 10-3 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 is a straight line, as shown by curve *A*. Such a modulation characteristic is perfectly **linear.**

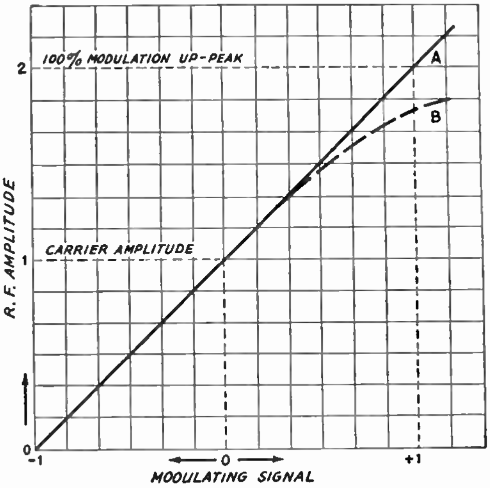


Fig. 10-3 — The modulation characteristic shows the relationship between the instantaneous amplitude of the r.f. output and the instantaneous amplitude of the modulating signal. The ideal characteristic is a straight line, as shown by curve *A*.

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.

The **modulation capability** of the transmitter is the maximum percentage of modulation that is possible without objectionable distortion from nonlinearity. The maximum capability can never exceed 100 per cent on the down-peak, but it is possible for it to be higher on the up-peak. The modulation capability should be as close to 100 per cent as possible, so that the most effective signal can be transmitted for a given carrier power.

Methods of 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. 10-4 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.

Audio Power

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.

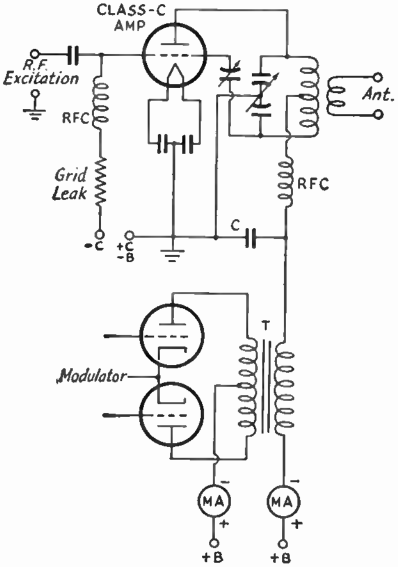


Fig. 10-4 — Plate modulation of a Class C r.f. amplifier. The r.f. plate by-pass condenser, *C*, in the amplifier stage should have reasonably high reactance at audio frequencies. (See chapter on modulators.)

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 the chapter on transmitters. 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 the chapter on 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. 10-5. The dropping resistor, R, should be of the proper value to apply normal d.c. voltage to the screen under steady carrier conditions. Its value can be calculated by taking the difference between plate and screen voltages and dividing it by the rated screen current.

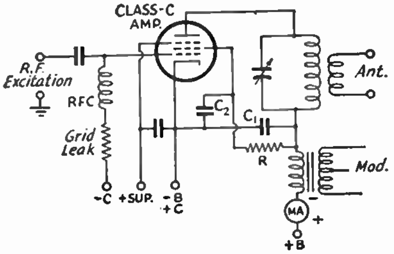


Fig. 10-5 — **Plate and screen modulation of a Class C r.f. amplifier using a pentode tube. The plate r.f. by-pass condenser, C_1, should have reasonably high reactance at all audio frequencies. (See chapter on modulators.) The screen by-pass, C_2, should be 0.002 μfd. or less in the usual case.**

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. 10-6. 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.

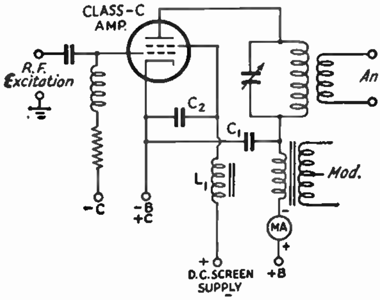


Fig. 10-6 — **Plate modulation of a beam tetrode, using an audio impedance in the screen circuit. The value of L_1 is discussed in the text. See Fig. 9-5 for data on by-pass capacitors C_1 and C_2.**

GRID MODULATION

The principal disadvantage of plate modulation is that a considerable amount of audio power is required. This can be avoided by applying the modulating signal to a grid element in the modulated amplifier, in which case the audio power needed is generally quite small.

In such systems 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 modulating 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 efficiency obtainable at the peak is of the order of 66 per cent, so the carrier efficiency ordinarily cannot exceed 33 per cent. For a given r.f. tube, the carrier output is about one-fourth the power obtainable from the same tube in c.w. operation.

The modulator is required to furnish only the audio power dissipated in the modulated grid under the operating conditions chosen. A speech amplifier capable of delivering 3 to 10 watts is usually sufficient.

Plate-Circuit Operating Conditions

The d.c. plate power 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. It is generally best to use the maximum plate voltage permitted by the manufacturer's ratings, because the optimum operating conditions are more easily achieved with high plate voltage and the linearity also is improved.

Example: Two tubes having plate dissipation ratings of 55 watts each are to be used with grid modulation.

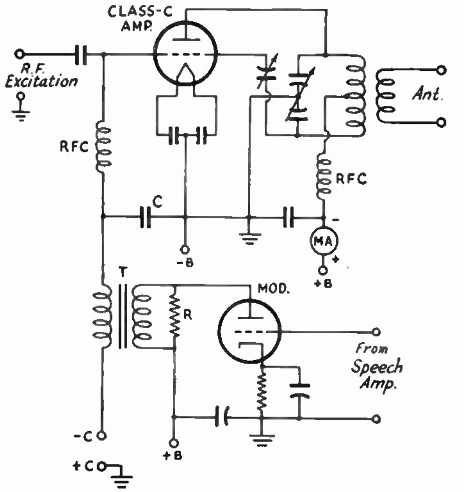


Fig. 10-7 — Control-grid modulation of a Class C amplifier. The r.f. grid by-pass condenser, *C*, should have high reactance at audio frequencies (0.005 μfd. or less).

The maximum permissible 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 average plate current for the two tubes will be

$$I = \frac{P}{E} = \frac{165}{1500} = 0.11 \text{ amp.} = 110 \text{ ma.}$$

At 33% efficiency, the carrier output to be expected is 55 watts.

The plate-voltage/plate-current ratio at *twice* carrier plate current is

$$\frac{1500}{220} = 6.8$$

The tank-circuit *L/C* ratio should be chosen on the basis of *twice* the average or carrier plate current. If the *L/C* ratio is based on the plate voltage/plate current ratio under carrier conditions the *Q* may be too low for good coupling to the output circuit.

Control-Grid Modulation

Control-grid modulation may be used with any type of r.f. amplifier tube. A typical triode circuit is given in Fig. 10-7. The same circuit can be used with screen-grid tubes merely by supplying the normal value of screen voltage by any convenient means; however, the screen should be by-passed for audio as well as radio frequencies. The audio signal is inserted, by means of transformer *T*, in series with the grid-bias lead. In a push-pull amplifier the transformer is connected in the common bias lead.

In control-grid modulation the d.c. grid bias is the same as in normal Class-C amplifier service, but the r.f. grid excitation is somewhat smaller. The audio voltage superimposed on the d.c. bias changes the instantaneous grid bias at an audio rate, thus varying the operating conditions in the grid circuit and controlling the output and efficiency of the amplifier.

The change in instantaneous bias voltage with modulation causes the rectified grid current of the amplifier also to vary, which places a variable load on the modulator. To reduce distortion, resistor *R* in Fig. 10-7 is connected in the output circuit of the modulator as a constant load, so that the overall load variations will be minimized. This resistor should be equal to or somewhat higher than the load into which the modulator tube is rated to work at normal audio output. The turns ratio of transformer *T* should be about 1 to 1 in most cases.

The load on the r.f. driving stage also varies with modulation. This in turn will cause the excitation voltage to vary which may cause the modulation characteristic to be nonlinear. To overcome it, the driver should be capable of two or three times the r.f. power output actually required to drive the amplifier. The excess power may be dissipated in a dummy load that then performs the same function in the r.f. circuit that resistor *R* does in the audio circuit.

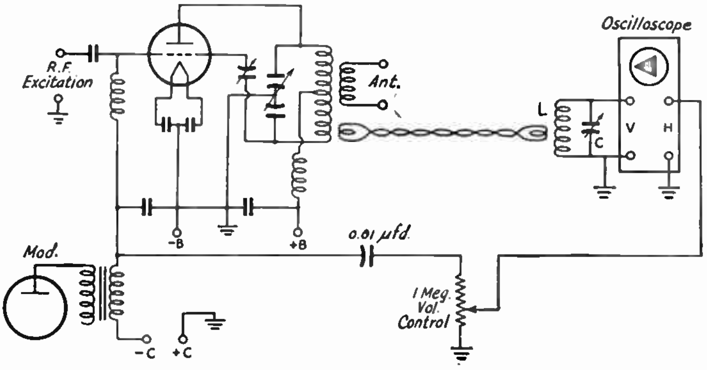
The d.c. bias source in this system should have low internal resistance. Batteries or a voltage-regulated supply are suitable. Grid-leak bias should not be used.

Adjustment

A control-grid modulated amplifier should be adjusted with the aid of an oscilloscope connected as shown in Fig. 10-8. 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

Fig. 10-8 — 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, 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.

The same set-up may be used for screen-grid and suppressor modulation, so long as the audio signal to the oscilloscope is coupled from the "hot" end of the coupling transformer secondary.



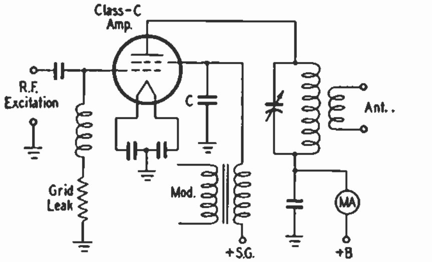


Fig. 10-9 — Screen-grid modulation of beam tetrode. Condenser *C* is an r.f. by-pass condenser and should have high reactance at audio frequencies. A value of 0.002 μfd. is satisfactory.

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 as linear as possible from the horizontal axis to twice the carrier amplitude.

Although the linearity of control-grid modulation is never as good as with properly-adjusted plate modulation, a satisfactory modulation characteristic can be obtained by using high plate voltage, high values of d.c. grid bias, and by careful adjustment, with the oscilloscope, of plate loading and r.f. grid excitation.

Screen-Grid Modulation

A typical circuit for modulating the screen grid of a beam tetrode power tube is given in Fig. 10-9. The modulated amplifier is operated Class-C, and the power output is varied at an audio-frequency rate by superimposing an audio voltage on the d.c. screen voltage. This can be done through a transformer as indicated in the diagram.

The peak audio voltage required for full modulation is slightly more than the operating value of d.c. screen voltage. The latter, in turn, is approximately half the rated screen voltage under maximum ratings for c.w. operation. The audio power required is approximately one-fourth the d.c. power input to the screen under c.w. operation.

In this method the amplifier is first adjusted to give optimum output with full rated screen voltage. The plate current and r.f. current into the antenna or feeders should be noted, and the d.c. screen voltage should then be reduced to one-half the initial value. If the plate current and r.f. output current both drop to one-half their initial values, the amplifier is properly adjusted. If not, the plate loading and grid excitation should be varied until this condition is approximated as closely as possible. When modulation is applied, using the reduced value of d.c. screen voltage, the amplifier can be modulated 100 percent.

The linearity with this method is generally good with beam tetrodes, and is better than can be obtained with control-grid modulation. The adjustments are also considerably less critical, particularly with respect to excitation power and control-grid bias. The normal grid-leak value is satisfactory, and best linearity usually is obtained with approximately half the d.c. grid current that is recommended for c.w. operation.

With pentodes, it is more difficult to obtain good linearity. Suppressor-grid modulation usually is to be preferred with such tubes.

An alternative method of modulating the screen grid of a beam tube is to use a "clamp" or screen-protective tube (see chapter on transmitters) as a Class-A modulator. Details of this method are given in March, 1950, *QST*, page 46.

Suppressor Modulation

The circuit arrangement for suppressor-grid modulation of a pentode tube is shown in Fig. 10-10. With tubes having suitable suppressor-grid characteristics, linear modulation up to practically 100 per cent can be obtained with negligible distortion.

The method of adjustment closely resembles that used with screen-grid modulation. The amplifier is first adjusted for optimum c.w. output with zero bias on the suppressor grid. Negative bias is then applied to the suppressor and increased in value until the plate current and r.f. output current drop to half their original values. When this condition has been obtained the amplifier is ready for modulation. Since the suppressor is always negatively biased, the modulator is not required to furnish any power.

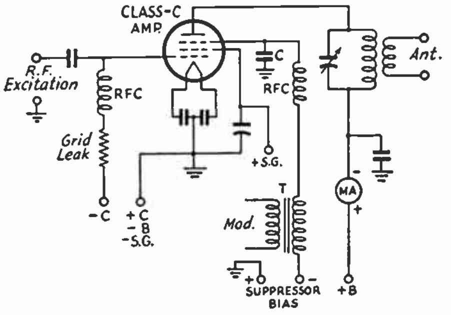


Fig. 10-10 — Suppressor-grid modulation of an r.f. amplifier using a pentode-type tube. The suppressor-grid r.f. by-pass condenser, *C*, should be the same as the grid by-pass condenser in control-grid modulation.

CATHODE MODULATION

Circuit

The fundamental circuit for cathode or "center-tap" modulation is shown in Fig. 10-11. This type of modulation is a combination of the plate and grid 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 control-grid 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. 10-12. 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 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 modulation, is at the right (*A*); pure grid modulation is represented by the left-hand ordinate (*B* and *C*).

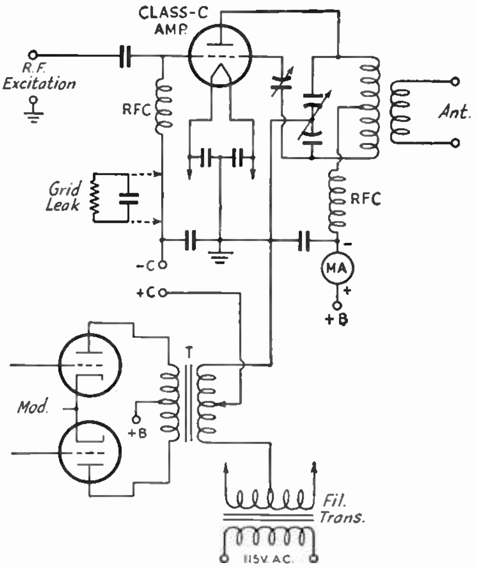


Fig. 10-11 — **Circuit arrangement for cathode modulation of a Class C r.f. amplifier.**

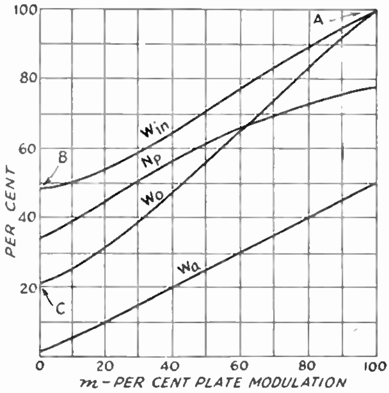


Fig. 10-12 — **Cathode-modulation performance curves, in terms of percentage of plate modulation plotted against percentage of Class C telephony tube ratings.**

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.

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. 10-12, 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 × 0.65 = 162.5 watts
Power output = 190 × 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 × 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 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.

● OTHER METHODS

It will be observed that in the methods of modulation so far described there are two basic ways of obtaining the necessary sideband power: 1) by supplying it in the form of audio-frequency power, as in plate modulation; 2) by obtaining it directly at r.f. through reduced carrier efficiency, as in grid modulation. Modulation systems are periodically proposed having the objective of avoiding both these conditions; i.e., the aim is to generate the sideband power without large amounts of audio power and without reducing the efficiency of the modulated amplifier at the carrier level.

For the most part such systems employ one amplifier operating at normal Class-C efficiency (about 70%) to supply the carrier. A second amplifier is used to aid the carrier tube in supplying the modulation peak. The second tube is usually idle with no modulation or during a modulation trough. Since the second tube also works at Class-C efficiency, the overall efficiency is high.

Modulation methods of this class are attractive economically, but are usually considerably more difficult to adjust than the systems described previously. They are also inherently incapable of giving as good linearity as the simpler plate- and grid-modulated systems. They have, as a result, had only sporadic use in amateur communication.

Controlled-Carrier Systems

With speech, the maximum modulation occurs during only part of the time because of the varying voice level, pauses between sentences, etc. It is possible to take advantage of this characteristic to increase the output of a grid-modulated system.

The method involves varying the carrier amplitude at a syllabic or somewhat slower rate so that the power input is low when there is little or no modulation. This reduces the plate dissipation at the time when the efficiency is least. Space does not permit a detailed treatment here, but information on practical systems can be found in the following *QST* articles:

Lippert, G. R., "A Constant-Modulation 'Phone System," *QST*, April, 1950.

Lippert, G. R., "Constant Modulation of the 813", *QST*, November, 1950.

Checking AM 'Phone Operation

Proper adjustment of a 'phone transmitter is aided immeasurably by the oscilloscope. The 'scope 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.

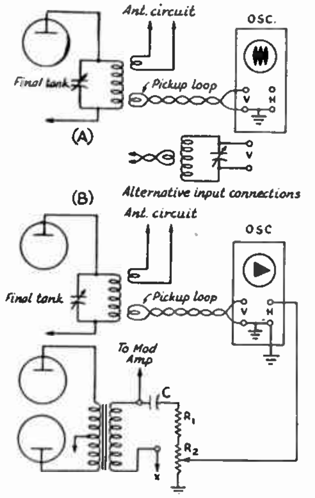


Fig. 10-13—Methods of connecting an oscilloscope to the modulated r.f. amplifier for checking modulation. See text for discussion.

Oscilloscope connections for both types of patterns are shown in Fig. 10-13. 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, 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. 10-14B, 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. 10-14D, 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. 10-13B. The vertical plates are coupled to the transmitter tank circuit 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

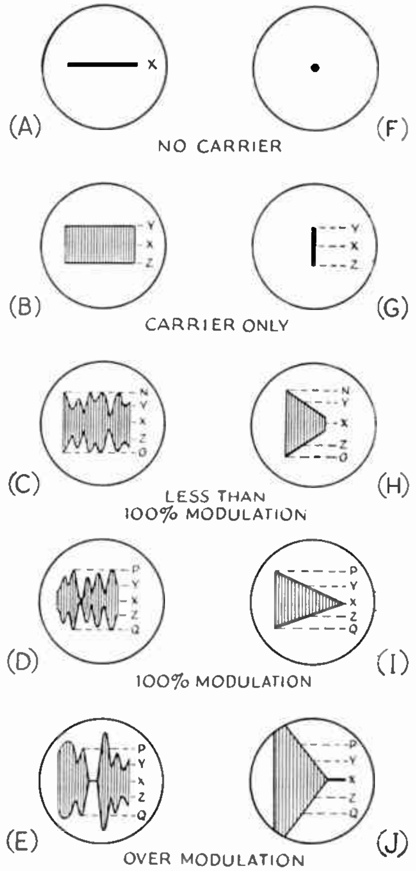
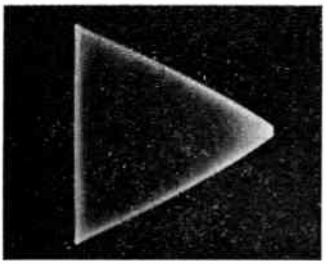
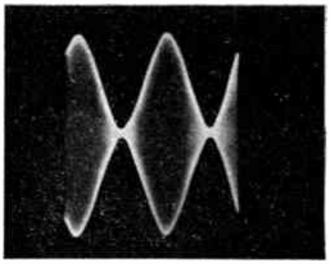
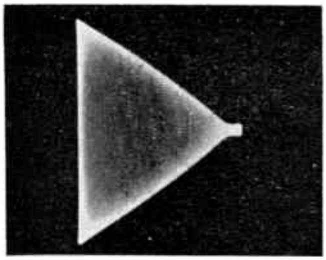


Fig. 10-14 — Wave-envelope and trapezoidal patterns representing different conditions of modulation.

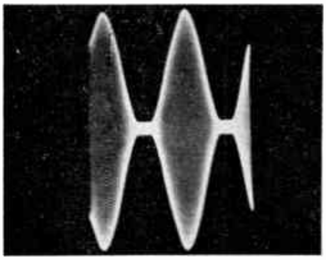


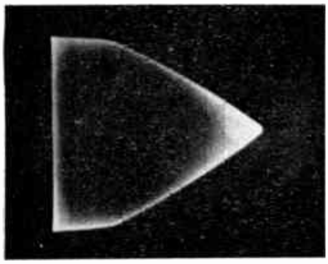
Properly-operated 'phone transmitter modulated 100 per cent.



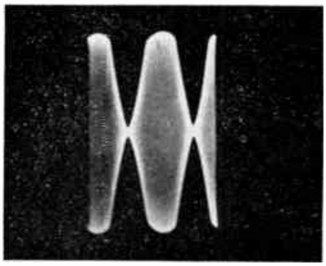


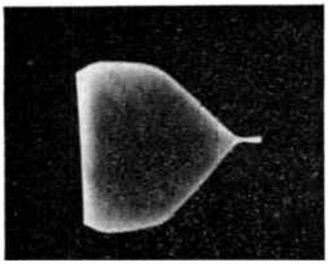
Overmodulation of a transmitter having high modulation capability. Distortion occurs only on the down-peaks.



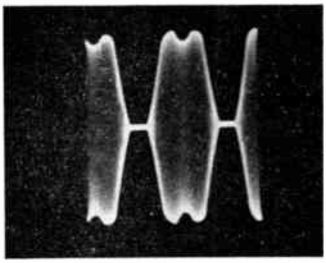


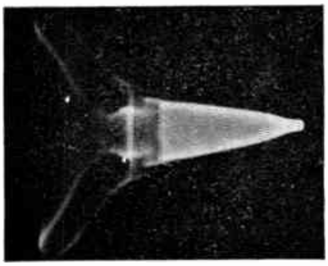
Nonlinearity in modulated r.f. stage, frequently caused by insufficient excitation of a plate-modulated amplifier or overexcitation of a grid-bias modulated amplifier. The amplifier modulates linearly in the downward direction but the up-peaks are flattened.



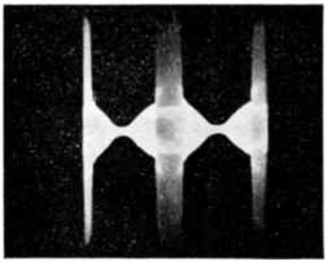


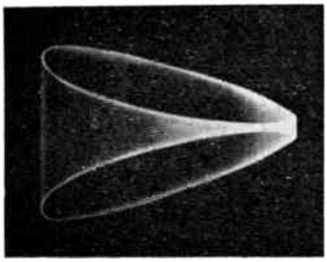
Overmodulation and nonlinear operation (insufficient modulation capability). These patterns are similar to those directly above, but with the modulation carried beyond 100 per cent in the downward direction.



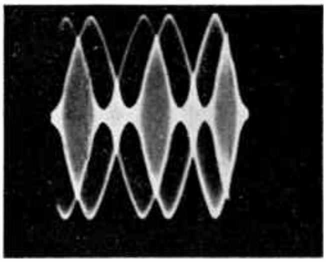


Overmodulation and parasitic oscillations in the modulated amplifier. The trapezoidal pattern also shows phase distortion caused by incorrect coupling between the oscilloscope and audio system.





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.)

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. In the example above, where *R* is 1 megohm, the capacitance should be at least 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 adequate voltage rating but insufficient capacitance are available.

Trapezoidal patterns for various conditions of modulation are shown in Fig. 10-14 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 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. 10-13B). 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. 10-14F). 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.

Curvature near the point of the wedge is generally to be expected with control-grid and suppressor modulation, because these methods cannot be made perfectly linear. With control-grid modulation the linearity is improved by operating with as much negative d.c. bias as possible, adjusting the r.f. excitation and plate loading for the best pattern.

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 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. 10-15 shows typical patterns of both the trapezoid and wave-envelope types. The cause of the distortion is indicated for grid-modulation systems. The patterns at A, although not truly linear, are representative of properly-operated control-grid modulation.

Faulty Patterns

The drawings of Figs. 10-14 and 10-15 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, 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 μμfd.) 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.

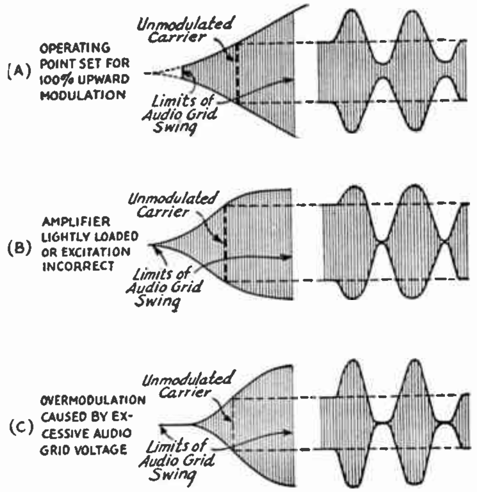


Fig. 10-15 — Oscilloscope patterns representing proper and improper adjustments for grid-modulation systems. Trapezoidal pattern at left; wave-envelope pattern at right. The pattern obtained with a correctly-adjusted amplifier is shown at A. The other drawings indicate nonlinear modulation from typical causes.

"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. 10-13B.

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.

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) The r.f. amplifier is not loaded properly to present the required value of modulating impedance to the modulator.
4) Insufficient output capacitance in the filter of the modulated-amplifier plate supply.
5) D.c. input to the r.f. amplifier, under carrier conditions, is in excess of the manufacturer's ratings for plate modulation. Alternatively, the filament emission of the amplifier tubes may be low.
6) In plate-and-screen modulation of tetrodes or pentodes, the screen is not being sufficiently modulated along with the plate. In systems in which the d.c. screen voltage is obtained through a dropping resistor, a downward dip in plate current may occur if the screen by-pass condenser capacitance is large enough to by-pass audio frequencies.

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 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. 10-15B; the pattern with an upward kick will look like Fig. 10-15A, 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 (to a cold water pipe, for example) 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.

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, compared with the carrier, 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 in the chapter on audio amplifiers, 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.

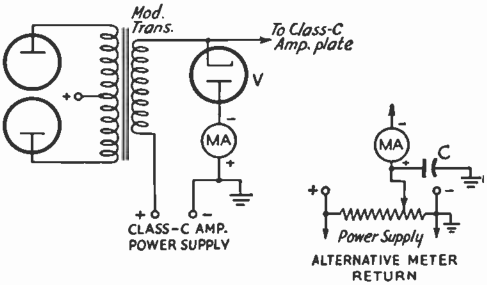


Fig. 10-16 — A negative-peak overmodulation indicator. Milliammeter *MA* may be any low-range instrument (up to 0–50 ma. or so). The inverse-peak-voltage rating of the rectifier, *V*, must be at least equal to the d.c. voltage applied to the plate of the r.f. amplifier. The alternative meter-return circuit 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.

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. 10-16 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 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 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.

CHAPTER 11

Frequency and Phase Modulation

Although the most common type of modulation is that in which the amplitude of the carrier is varied, it is also possible to convey intelligence by varying the frequency or phase of the carrier.

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.

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. 11-1 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

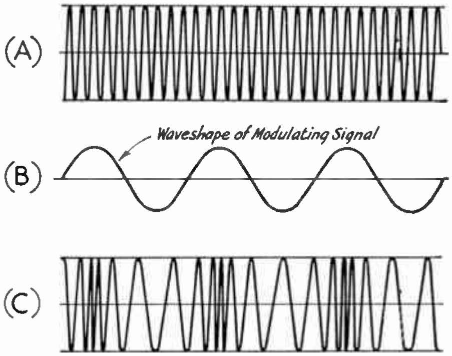


Fig. 11-1 — Graphical representation of frequency modulation. In the unmodulated carrier at A, each r.f. cycle occupies the same amount of time. When the modulating signal, B, is applied, the radio frequency is increased and decreased according to the amplitude and polarity of the modulating signal.

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,

$$\textit{Modulation index} = \frac{\textit{Carrier frequency deviation}}{\textit{Modulating frequency}}$$

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. 11-2 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. 11-2, 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

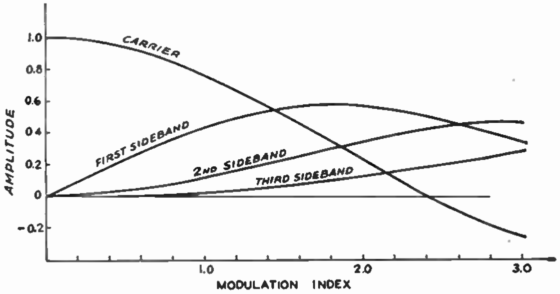


Fig. 11-2 — How the amplitude of the pairs of sidebands varies with the modulation index in an FM or PM signal. If the curves were extended for greater values of modulation index it would be seen that the carrier amplitude goes through zero at several points. The same statement also applies to the sidebands.

plate modulation, and by changing the power input and efficiency in the case of grid-bias modulation.

The curves of Fig. 11-2 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. 11-2 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. 11-2, 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 carrier power.

Comparison of FM and PM

The methods used by amateurs for the reception of FM or PM signals (see receiving chapter) 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 the receiving chapter 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. 11-3 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. 11-4. 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. 11-3; it represents the input capacitance of the tube. (It is possible, also, to omit C_1 from Fig. 11-3 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. 11-4 to permit controlling the sensitivity.) In Fig. 11-4 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. 11-3.

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.

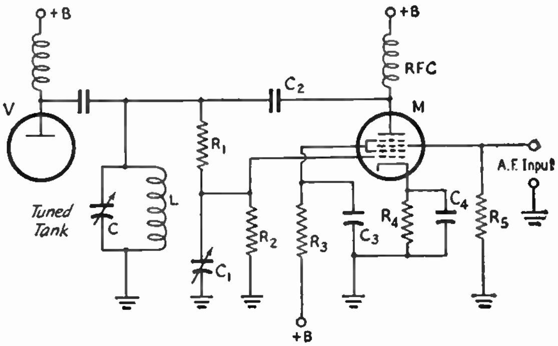


Fig. 11-3 — Reactance-modulator circuit using a 6L7 tube.
C — R.f. tank capacitance. C_1 — 3-30 $\mu\mu$fd. C_2 — 220 $\mu\mu$fd.
C_3 — 8-μfd. electrolytic (a.f. by-pass) in parallel with 0.01-μfd. paper (r.f. by-pass).
C_4 — 10-μfd. electrolytic in parallel with 0.01-μfd. paper.
L — R.f. tank inductance.
R_1 — 47,000 ohms.
R_3 — 33,000 ohms.
R_2, R_5 — 0.47 megohm.
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.

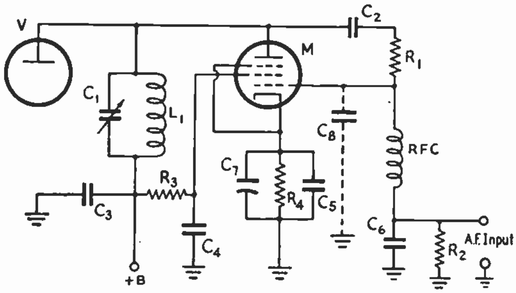


Fig. 11-4 — Reactance modulator using a high-transconductance pentode (6SG7, 6AG7, etc.).

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. 11-3 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. 11-3 and 11-4 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. 11-5 and 11-6 uses a pentode reactance modulator in a circuit which is basically that of Fig. 11-4. 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. 11-4. 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. 11-5 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.

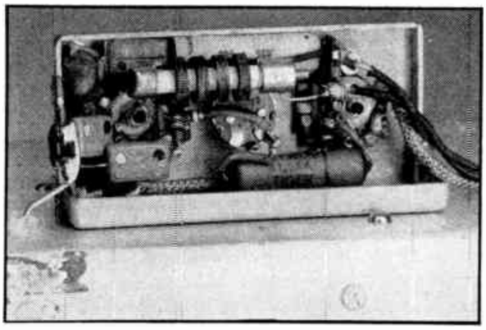


Fig. 11-6 — **Underneath the modulator unit. The r.f. connection to the VFO goes through the feed-through bushing at the left.**

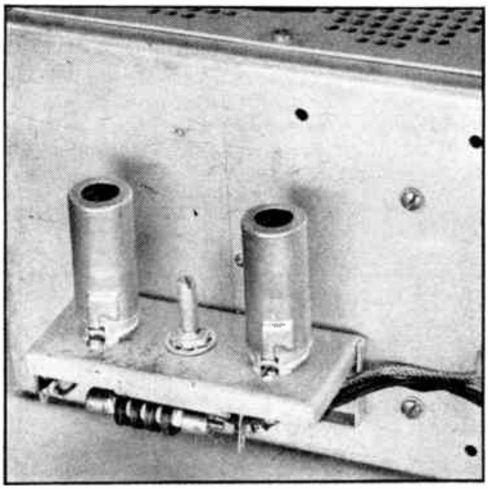


Fig. 11-5 — **Miniature reactance modulator that can be used with any VFO. The shielded lead is for microphone input; the other two wires bring in filament and plate supply.**

The adjustment of reactance modulators is discussed in a later section in this chapter.

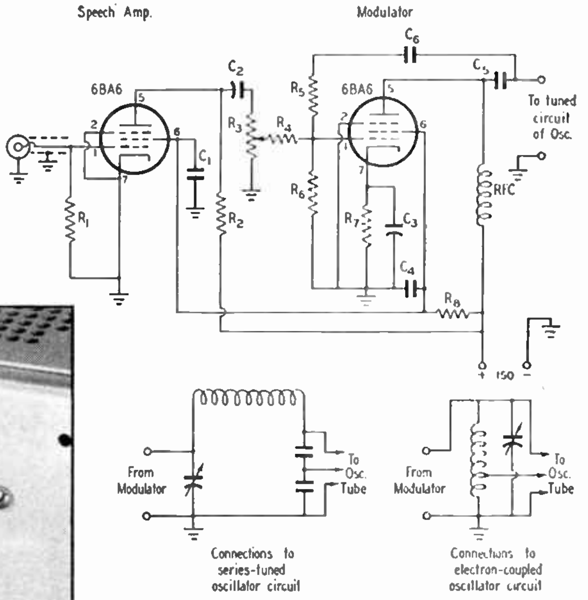


Fig. 11-7 — **Circuit diagram of the narrow-band FM modulator unit.**

C_1 — 680-μμfd. mica.
C_2, C_4 — 0.01-μfd. paper, 400 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.

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 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.

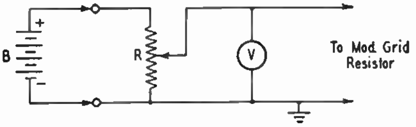


Fig. 11-8 — D.c. method of checking frequency deviation of a reactance-tube-modulated oscillator. A 500- or 1000-ohm potentiometer may be used at *R*.

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. 11-8. 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 miscellaneous data chapter 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. 11-9. 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. 11-10. Note its deflection (using the d.c. voltage method as in

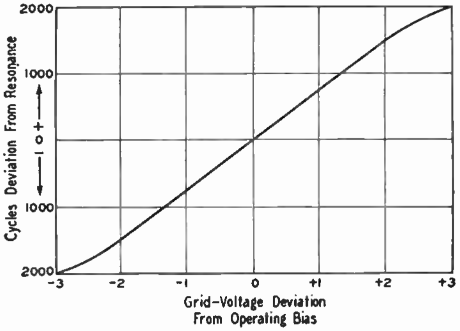


Fig. 11-9 — A typical curve of frequency deviation *vs.* modulator grid voltage.

Fig. 11-8) 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.

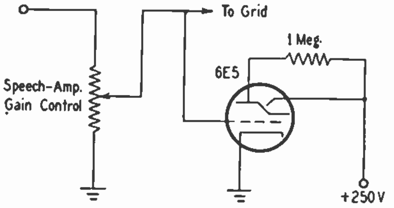


Fig. 11-10 — 6E5 modulation indicator for FM or PM modulators. To insure sufficient grid voltage for a good deflection, it may be necessary to connect the gain control in the modulator grid circuit rather than in an earlier speech-amplifier stage.

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 (through an attenuator, if necessary, to avoid overloading; see chapter on audio amplifiers) and increase the audio gain until there is a small amount of modulation. Tuning the receiver near 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.

CHAPTER 12

Single-Sideband Transmitting Techniques

The most significant development in amateur radiotelephony in the past several years has been the increased use of single-sideband suppressed-carrier transmissions. 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.

● SUPPRESSING THE CARRIER

The carrier can be suppressed or nearly eliminated by an extremely sharp filter or by using a **balanced modulator**. The basic principle in any balanced modulator is to introduce the carrier in such a way that it does not appear in the output but so that the sidebands will. This requirement is satisfied by introducing the audio in push-pull and the r.f. drive in parallel, and connecting the output (plate circuit) of the tubes in push-pull. Balanced modulators can also be connected with the r.f. drive and audio inputs in push-pull and the output in parallel, with equal effectiveness. The choice of a balanced modulator circuit is generally determined by constructional considerations and the method of modulation preferred by the builder.

In any of the circuits, there will be no output with no audio signal because the circuits are balanced. The signal from one tube is balanced or cancelled in the output circuit by the signal from the other tube. The circuits are thus balanced for any value of *parallel* audio signal. When push-pull audio is applied, the modulating voltages are of opposite polarity, and one tube will conduct more than the other. Since any modulation process is the same as "mixing" in receivers, sum and difference frequencies (sidebands) will be generated. The modulator is not balanced for these sidebands, and they will appear in the output.

The amount of carrier suppression is dependent upon the matching of the two tubes and their associated circuits. Normally two tubes of the same type will balance closely enough to give at least 15 or 20 db. carrier suppression without any adjustment. If further suppression is required, trimmer condensers to balance the grid-plate capacities and bias adjustments for setting the operating points can be used.

● SINGLE-SIDEBAND GENERATORS

Two basic systems for generating SSB signals are shown in Fig. 12-1. 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 500 kc. by using multiple-crystal filters. The low-frequency oscillator output is combined with the audio output

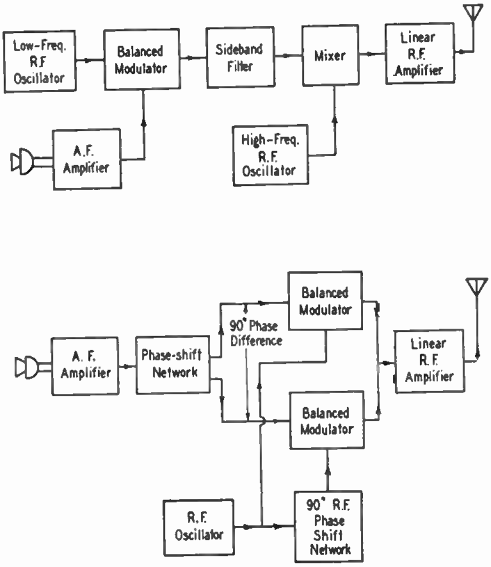


Fig. 12-1 — Two basic systems for generating single-sideband suppressed-carrier signals.

of a speech amplifier in a balanced modulator, 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 problem of image frequencies in the frequency conversions of SSB signals differs from the problem in receivers because the beating-oscillator frequency becomes important. Either balanced modulators or sufficient selectivity must be used to eliminate the possibility of unwanted radiations.

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 following amplifier.

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 one frequency conversion after modulation. The phasing system requires fewer stages and can be designed to require no frequency conversion, 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.

Regardless of the method used to generate a SSB signal, the minimum cost will be found to be higher than for an AM transmitter of the same low power. However, as the power level is increased, the SSB transmitter becomes more economical than the AM rig, both basically and from an operating standpoint.

● 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. 6-18 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 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. 12-2. 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.

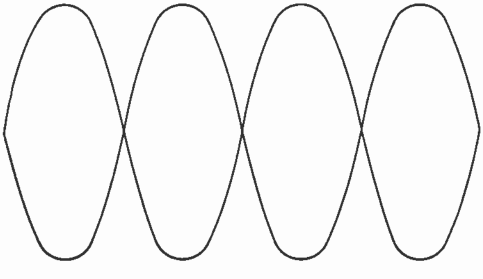


Fig. 12-2 — Oscilloscope pattern obtained with a two-tone test signal through a correctly-adjusted linear amplifier.

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-side-band 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. 12-3. 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.

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.

● VOICE-CONTROLLED BREAK-IN

Although it is possible for two SSB stations operating on widely different frequencies to work "duplex" if the carrier suppression is great enough (inadequate carrier suppression would be a violation of the FCC rules), most SSB operators prefer to use voice-controlled break-in and operate on the same frequency. This overcomes any possibility of violating the FCC rules and permits three or more stations to engage in a "round table." Voice-controlled break-in is not popular with straight AM because turning the carrier on and off at a syllabic rate results in a "keyed" type of heterodyne interference that is particularly annoying.

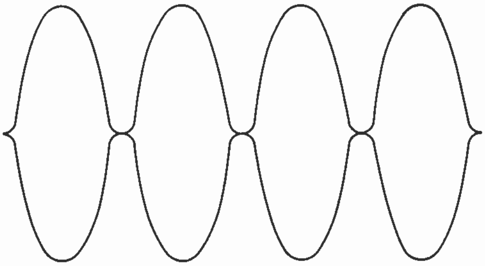


Fig. 12-3 — The distorted two-tone test-signal pattern obtained when the bias voltage is incorrect.

Many various systems of voice-controlled break-in are in use, but they are all basically the same. Some of the audio from the speech amplifier is amplified and rectified, and the resultant d.c. signal is used to key an oscillator and one or more stages in the SSB transmitter and "blank" the receiver at the time that the transmitter is on. Thus the transmitter is on at any and all times that the operator is speaking but is off during the intervals between sentences. The voice-control circuit must have a small amount of "hold" built into it, so that it will hold in between words, but it should be made to turn on rapidly at the slightest voice signal coming through the speech amplifier. Both tube and relay keyers have been used with good success.

A Phasing-Type SSB Exciter

The exciter shown in Figs. 12-4, 12-6 and 12-8 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 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 schematic of the exciter is shown in Fig. 12-5. 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.

Each balanced-modulator tube grid is fed through a blocking condenser and provided with grid-leak bias. The bias circuit of each balanced modulator is made adjustable for control of the carrier suppression. Provision is also made for the addition of fixed bias, in case the exciter is used in a voice-controlled circuit where the r.f. excitation is removed during listening periods.

Screen modulation is used, and the screen of each modulator tube is by-passed to ground for r.f. A transformer with a center-tapped secondary is used in the output of each audio amplifier to provide push-pull modulating voltages.

A reversing switch, S_1, allows switching to either the upper or lower sideband. If this switch has a center "off" position, it will facilitate using the "two-tone test" procedure mentioned earlier. A voltage 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.

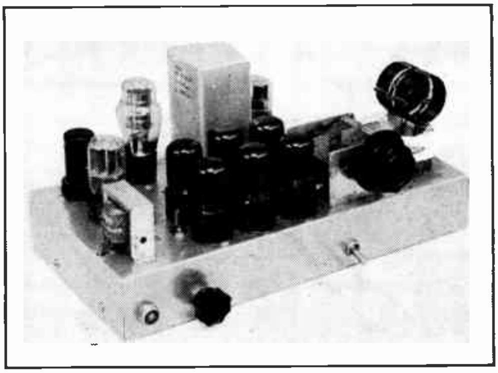


Fig. 12-4 — A small single-sideband exciter that includes voice-controlled break-in. Receiving-type tubes are used throughout.

Microphone input and audio gain control are at the left-hand side of the front — the switch selects the upper or lower sideband. (Revised version, W2UNJ, Aug., 1949, *QST.*)

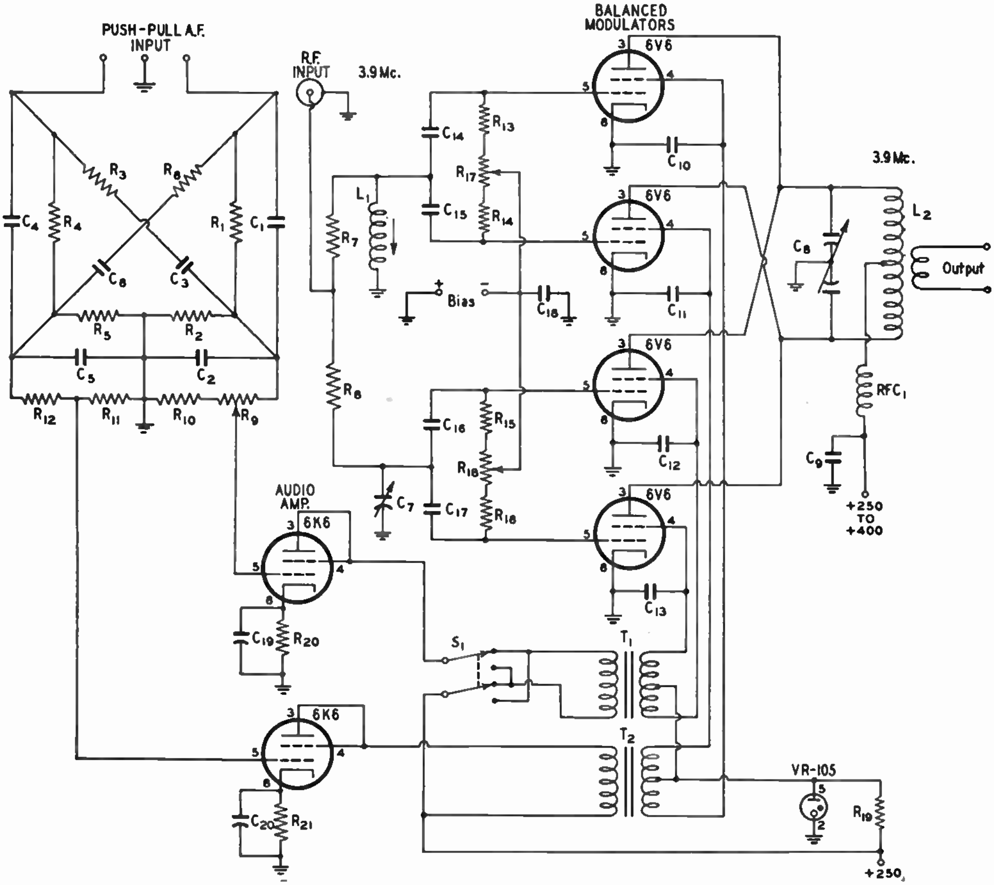


Fig. 12-5 — Circuit diagram of the single-sideband exciter.

C_1–C_6 — See Table 12-I.
C_7 — 150-μμfd. air padder condenser.
C_8 — Approx. 400-μμfd. per section, b.c. receiver tuning condenser.
C_9 — .001-μfd. 1000-volt mica.
C_{10}–C_{18} — .001-μfd. 500-volt mica.
C_{19}, C_{20} — 4-μfd. 150-volt electrolytic.
R_1–R_6 — See Table 12-I.
R_7, R_8 — 300 ohms, 5 watts (5 1500-ohm 1-watt in parallel).
R_9 — 0.5 megohm linear volume control.
R_{10} — 0.47 megohm.
R_{20} — 0.75 megohm.
R_{21} — 0.24 megohm.
R_{13}–R_{16} — 10,000 ohms.
R_{17}, R_{18} — 15,000-ohm potentiometer, wirewound.
R_{19} — 7500 ohms, 10 watts.
R_{20}, R_{21} — 680 ohms, 2 watts.
All resistors 1-watt unless specified otherwise.
L_1 — 25 turns No. 28 enam. closewound at mounting end of slot of National XR-50 slug-tuned form.
L_2 — 40-meter 75-watt tank coil with swinging link (Bud 6-OLS-40).
RFC_1 — 2.5-mh. r.f. choke.
S_1 — D.p.d.t. toggle, preferably with center off. See text.
T_1, T_2 — 5-watt modulation transformer, 10,000 ohms c.t. to 4000 ohms (Stancor A-3812).

Speech Amplifier and Voice Control

The speech amplifier is designed to attenuate both low and high frequencies, amplifying only the audio range required for good intelligibility. The wiring diagram is shown in Fig. 12-7. The output of the speech amplifier is coupled to the input of the audio phase-shift network through a transformer with a center-tapped secondary, to provide push-pull audio for the phase-shift network.

Part of the output of the speech amplifier is taken off through an adjustable voltage divider circuit and blocking condenser to the voice-control circuit. There it is rectified by the diodes of the 6SQ7, and the resulting d.c. voltage is used to charge C_{14} negative. An audio choke prevents audio components from appearing across C_{14}. The triode section of the 6SQ7 is normally conducting and holding the relay closed, but when the negative voltage appears across C_{14} the 6SQ7 plate current is cut off and the relay opens. When the audio signal is removed, C_{14} discharges through R_{15} and the triode again conducts, closing the relay.

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 12-I is used in selecting the network

TABLE 12-I
Phase-Shift Network Design Data

Part	Nominal Value	Target Value	Measured Value
C_1	0.001	0.00105	(Cm_1)
C_2	0.002	0.00210	(Cm_2)
C_3	0.006	0.00630	(Cm_3)
C_4	0.005	0.00475	(Cm_4)
C_5	0.01	0.00950	(Cm_5)
C_6	0.03	0.0285	(Cm_6)
R_1	100,000	$\frac{100}{Cm_1} =$	
R_2	50,000	$\frac{105}{Cm_2} =$	
R_3	15,000	$\frac{100}{Cm_3} =$	
R_4	100,000	$\frac{453}{Cm_4} =$	
R_5	50,000	$\frac{476}{Cm_5} =$	
R_6	15,000	$\frac{453}{Cm_6} =$	

All condensers mica, and all resistors 1 watt.

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.

Construction

The exciter and its associated audio equipment are assembled on a 13 by 17 by 2-inch aluminum chassis. The four 6V6 balanced-modulator tubes are arranged in a square pattern toward the front center of the chassis, with the plate tuning condenser and coil off to one side and the 6K6 audio amplifier tubes on the other. The two modulation transformers are under the chassis directly below the plate tuning condenser. The speech amplifier is arranged along the left-hand side of the chassis, with the 6SJ7 at the rear and the output transformer on the top of the chassis at the front. The audio phase-shift network is below the output transformer.

The reactive components of the r.f. phasing network, L_1 and C_7, are mounted in a plug-in shield can that mounts directly behind the balanced-modulator tubes. The shield can is grounded to the chassis through the spare pins of its plug. The voltage regulator tube is mounted to the left of the shield can, and the 6SQ7 voice-control tube is to the right. The components in the voice-control circuit are mounted under the chassis at the rear.

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.

The plate voltage for the speech amplifier must not be taken from the same point in the power supply that furnishes voltage for the 6K6 amplifiers, since interaction may occur that will upset the phase relationship at the output of the two 6K6s. If separate plate voltage sources are not available, an added filter section may be used to isolate the voltage to the speech amplifier.

The built-in voice-controlled relay can be used in a number of ways to provide the rapid voice break-in commonly used on 3.9-Mc. SSB 'phone. If a good c.w. break-in system is already in use at the station, the voice-control relay contacts may be substituted for the key, and no other changes are necessary.

If the local oscillator in the receiver will key in the plate voltage lead satisfactorily, then a simple voice break-in system may be obtained by using the relay contacts to shift the plate voltage from the receiver local oscillagor to the VFO. A drifting receiver oscillator must be avoided in this system, however.

Operating Conditions

If voice control is not used, and d.c. operating voltages are removed when excitation is removed for stand-by, then no fixed bias is required on the balanced modulators and a jumper can be placed

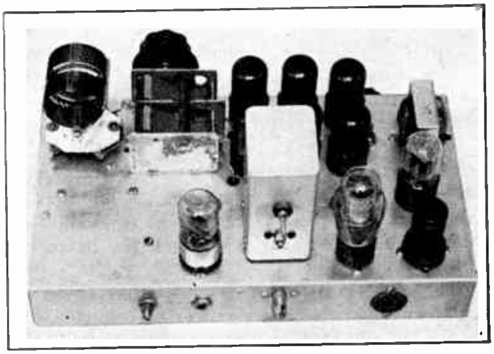


Fig. 12-6 — A rear view of the phasing-type exciter. The two r.f. phasing adjustments project from the shield can. The potentiometer shaft at the left sets the voice-control threshold level. The jack is for the keyed circuit, the r.f. connector takes the excitation cable, and the octal socket is for the power cable.

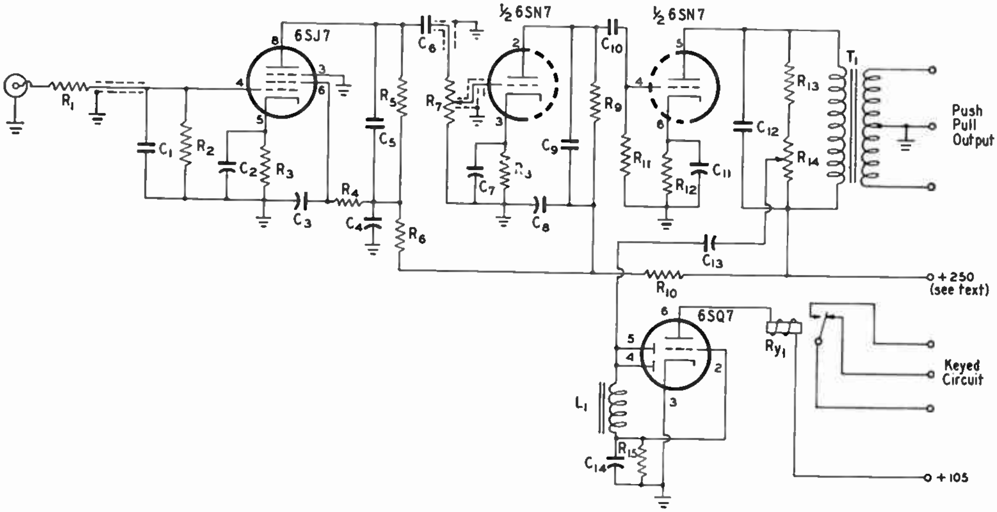


Fig. 12-7 — Wiring diagram of the speech amplifier and voice-control circuit.

C_1 — 100-μμfd. mica or ceramic.
C_2, C_7, C_{11} — 4-μfd. 150-volt electrolytic.
C_3 — .02-μfd. 400-volt paper.
C_4, C_8 — 8-μfd. 450-volt electrolytic.
C_5 — 270-μμfd. mica or ceramic.
C_6 — .001-μfd. mica or ceramic.
C_9 — .0033-μfd. mica or ceramic.
C_{10} — .002-μfd. mica or ceramic.
C_{12} — .005-μfd. ceramic or mica.
C_{13} — .01-μfd. 400-volt paper or ceramic.
C_{14} — .5-μfd. 200-volt paper. See text.
R_1, R_9 — 0.1 megohm.
R_2 — 2.2 megohm.
R_3, R_{12} — 910 ohms.
R_4 — 1.0 megohm.
R_5 — 0.27 megohm.
R_6 — 27,000 ohms.
R_7 — 0.5-megohm volume control.
R_8 — 2700 ohms.
R_{10}, R_{13} — 10,000 ohms, 1 watt.
R_{11}, R_{15} — 0.47 megohm.
R_{14} — 15,000-ohm volume control.
All resistors ½-watt unless specified otherwise.
T_1 — 5-watt modulation transformer, 10,000 ohms c.t. to 4000 ohms (Stancor A-3812).
L_1 — Small filter or audio choke (Stancor C-1707).
Ry_1 — Sensitive 10,000-ohm relay.

across the bias terminals. When excitation is removed with d.c. voltages applied, as in voice-controlled operation, then 4½ volts of fixed bias should be used to limit the plate and screen currents on the balanced modulators.

With 400 volts applied to the balanced-modulator plates and 250 volts to all other plate supply inputs, the operating currents will be approximately as follows:

Total balanced-modulator plate current	85 ma.
VR tube supply current	20 ma.
Total 6K6 amplifier current	62 ma.
Total speech amplifier current	12 ma.

The total balanced-modulator grid current, measured at the bias terminals, will vary with excitation, but it should be in the range 3 to 5 ma.

These currents will not change appreciably with varying audio input and, with the exception of the grid current, will not change appreciably when the excitation is removed, provided that 4½ volts of fixed bias is used on the balanced-modulator grids.

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, the grid tank circuit will require shunting 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.

Adjustment

Adjustment of the exciter is best made under actual operating conditions. Connect the exciter to the transmitter, load the exciter with a dummy load, apply r.f. excitation, feed sine-wave audio into the speech amplifier, and tune in the conventional way for maximum output.

Reduce the audio input to zero, and adjust potentiometers R_{26} and R_{27} 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.

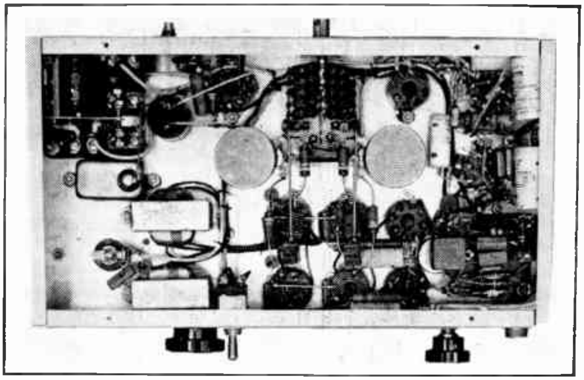


Fig. 12-8 — Underneath the chassis of the exciter. The two potentiometers are the bias balancing controls, R_{17} and R_{18}.

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_9 for 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 voltage on all four balanced modulator tubes will be approximately 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.

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. 12-9. 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. at 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.

The exciter is capable of driving any pair of beam tubes commonly used in amateur transmitters, or any pair of triodes in Class AB1. A buffer stage will ordinarily be required to drive Class-B triodes.

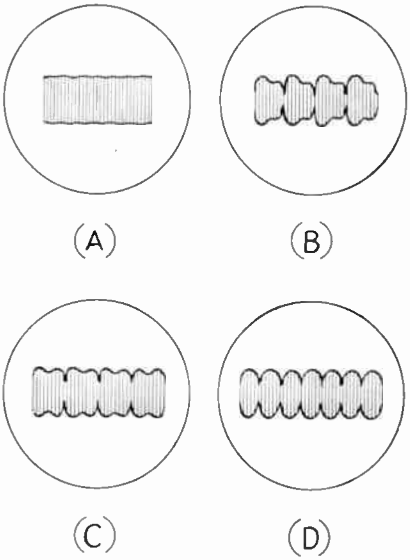


Fig. 12-9 — Sketches of the oscilloscope face showing different conditions of adjustment of the exciter unit. (A) shows the substantially clean carrier obtained when all adjustments are at optimum and a sine-wave signal is fed to the audio input. (B) shows improper r.f. phase and unbalance between the 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.

A Crystal-Filter SSB Exciter

The exciter uses a quartz crystal filter operating at 450 kc. (or vicinity). The filter allows a passband of 300 to 3000 cycles; the sideband rejection should run 35-40 db. over 300 to 3000 cycles. At no time within the reject range is the rejection less than 30 db.; at some places it approaches 60 db. Suppression of the carrier is obtained without the use of balanced modulators, and the stability of suppression is excellent. Crystals suitable for use in the filter are available on the war surplus market for less than one dollar each. The most useful of these crystals are in the series that runs from 375 to 525 kc. in 1.388-kc. steps; this series is marked at 72 times the crystal frequency in a series of channels from 28.0 to 38.0 Mc. The crystals were manufactured by Western Electric for the Signal Corps, and are of the plated variety, mounted in an FT-241A holder. The holder pins have ½-inch spacing. The crystals may be socket-mounted or soldered directly into the filter at the builder's discretion.

The filter is of bridge design with complex entry and terminating sections. The complex sections are used to suppress the carrier and modify the response characteristics of the bridge. Fig. 12-10 shows the filter proper, set for rejection of the upper sideband. The transformer, T_1, is a replacement-type 455-kc. inter-stage i.f. transformer, mica-tuned, and air-cored. T_2 is also a replacement type, designed to feed into a diode detector.

The original filter was designed to operate at a carrier frequency of 450 kc., although the filter will work at frequencies between 425 and 490 kc. without alteration of the circuit or transformers. Under the condition of design for 450-kc. carrier, crystal "B" is 2.78 kc. *higher* than 450 kc., or 2 channels higher in the crystal series. Crystal "C" is 1.39 kc. *lower* than 450 kc., or 1 channel lower. Crystal "D" is 450 kc. Crystal "A," also at 450 kc., is used in a crystal oscillator

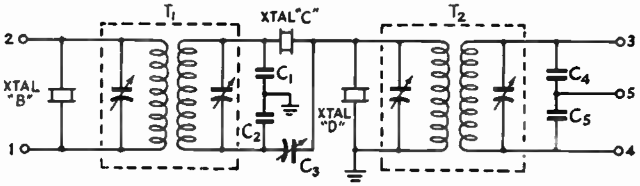


Fig. 12-10 — The 450-kc. quartz crystal filter used for sideband and carrier rejection.

C_1, C_2, C_4, C_5 — 100-μμfd. mica or ceramic.
C_3 — 3- to 30-μμfd. ceramic trimmer.
T_1 — 455-kc. interstage i.f. transformer (Meissner 16-6659).
T_2 — 455-kc. diode i.f. transformer (Meissner 16-6660).
For a carrier frequency of 450 kc., the crystals are:

Crystal	*B*	*C*	*D*
High-freq. reject	452.8 kc.	448.6 kc.	450.0 kc.
Low-freq. reject	447.2 kc.	451.4 kc.	450.0 kc.

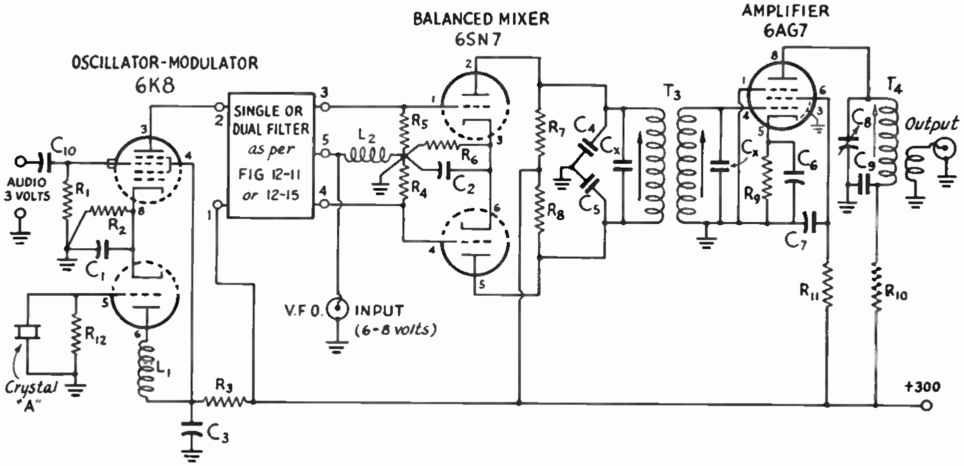


Fig. 12-11 — Complete diagram of the crystal-filter SSB exciter.

C_1, C_2, C_3, C_6, C_7 — 0.1-μfd. 400-volt paper.
C_4, C_5 — 39-μμfd. ceramic.
C_8 — 100-μμfd. variable air condenser.
C_9 — 0.02-μfd. 600-volt mica.
C_{10} — 0.01-μfd. 400-volt paper.
C_X — Trimmers in T_3.
R_1 — 0.47 megohm.
R_2 — 220 ohms.
R_3, R_{11} — 20,000 ohms, 1 watt.
R_4, R_5 — 0.1 megohm.
R_6, R_7, R_8 — 10,000 ohms.
R_9 — 150 ohms, 1 watt.
R_{10} — 1000 ohms.
R_{12} — 47,000 ohms.
All resistors ½ watt unless specified otherwise.
L_1 — 2.5-mh. r.f. choke.
L_2 — 0.5-mh. r.f. choke.
T_3 — 5-Mc. slug-tuned i.f. transformer.
T_4 — 5-Mc. slug-tuned i.f. transformer. Secondary removed and 8-turn link wound over cold end of primary. All fixed capacitors removed.

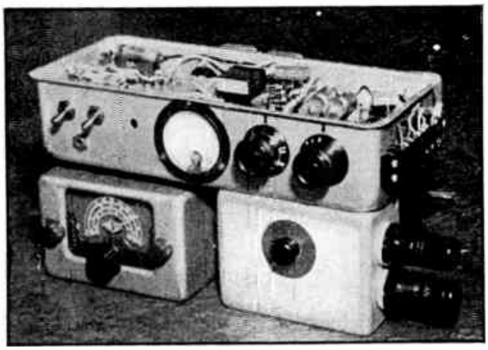


Fig. 12-12 — The crystal-filter SSB exciter, as designed for mobile work, complete with receiver converter and VFO. The top dish is the exciter (with cover removed). The meter reads cathode current to a pair of 807s driven by the unit, and the two knobs handle carrier reinsertion and 6AG7 plate tuning. (W1JEO/9, Nov., 1950, *QST*.)

to generate the initial carrier. Channel markings on these crystals are as follows:

"A" — 32.4 Mc., Channel 324
"B" — 32.6 Mc., Channel 326
"C" — 32.3 Mc., Channel 323
"D" — 32.4 Mc., Channel 324

Any other group within the range of the i.f. transformers may be utilized; only the channel relationship need be retained.

A diagram of the exciter proper is shown in Fig. 12-11. The 6K8 hexode-triode serves as 450-kc. oscillator and audio mixer. Approximately 3 volts of audio is required at the signal grid of the 6K8 for optimum results. The 6K8 delivers a carrier (450 kc.) and sidebands to the input of the filter. The filter rejects one sideband (depending upon the selection of crystals) and delivers single-sideband energy to the 6SN7 mixer. The filter also suppresses the carrier some 60 db. below the peak sideband energy. The 6SN7 mixer combines the single-sideband energy (in the vicinity of 450 kc.) with the output of the VFO (3400 to 3550 kc.) and the sum products are recovered in the output (3850 to 4000 kc.). The balanced mixer is used to remove the VFO component from the output tank. Balance is not critical and no adjustments are required or provided. A VFO signal of about 6 to 8 volts is required. The output of the mixer is fed to the grid of a 6AG7 which runs as a Class A tuned r.f. amplifier. The output of the 6AG7 is sufficient to drive a pair of 807s Class AB_2. Operation on 10 and 20 meters can be accomplished by heterodyning again to the desired band. Most VFOs in use cover or may be easily made to cover 3400 to 3550 kc. A single untuned 6SJ7 or 6AC7 Class A amplifier following a BC-221 might be used as a driver for this exciter.

Construction

The original transmitter was built for mobile operation and much hole drilling and experimentation has occurred on the chassis. Mounting the crystals on opposite sides of the transformers will keep stray capacity coupling at a minimum. No shielding other than that provided by the i.f. cans and the output tank can is required. It is important that capacity coupling *around* the crystal filter be minimized — in other words, no modulated signal must reach the 6SN7 mixer by any route except through the filter. Before construction is started, a decision must be made as to whether or not choice of sidebands is desired. If choice of sidebands *is* desired, a dual filter using 5 crystals will be required. This filter is shown schematically in Fig. 12-13. A double-section wafer switch selects the upper or lower sideband. These wafer sections must be separated by approximately 3 inches to minimize stray coupling. It is recommended that the crystals be wrapped with several layers of adhesive tape and then strapped to the chassis with metal brackets; connections may then be made by soldering to the holder pins.

Alignment

Alignment of the filter is straightforward, and once aligned it will need little attention.

1) Crystal "A" is first removed from the circuit. This crystal is best provided with a socket

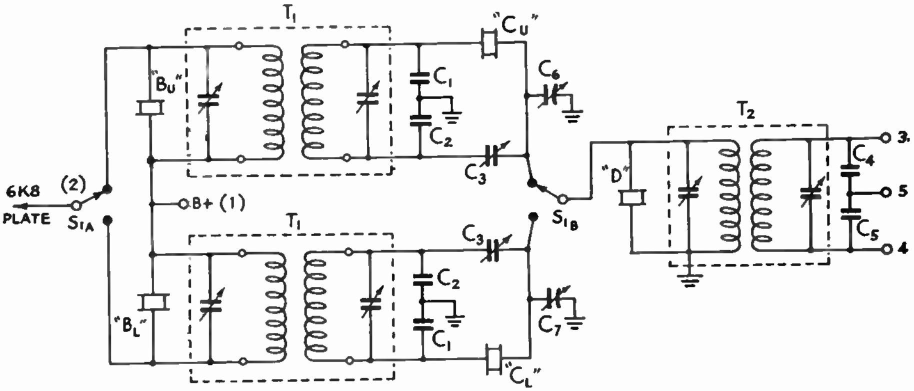


Fig. 12-13 — The double-channel crystal filter. All components are the same as in Fig. 12-10, except for the addition of the d.p.d.t. wafer switch, S_1, and the compensating condensers, C_6 and C_7 (3- to 30-μμfd. ceramic). The trimmer on the input side of T_2 is set at minimum and the alignment procedure is followed with C_6 or C_7 wherever the instructions call for adjusting the input condenser.

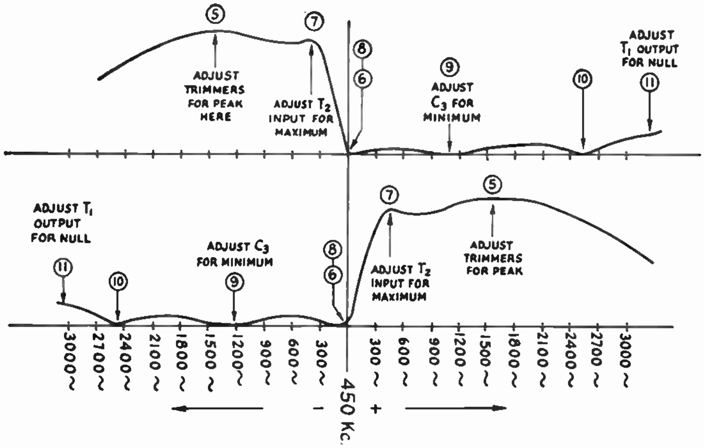


Fig. 12-14 — An alignment chart of the crystal filter. The numbers in the circles correspond to the steps outlined in the text.

mount so it can be removed during alignment.

2) A calibrated signal generator covering the crystal range is connected to the grid of the triode section of the 6K8.

3) A vacuum tube voltmeter is connected from grid to ground of one of the 6SN7 grids.

4) Swing the signal generator through the crystal range until a maximum response is noted at the voltmeter. This will indicate the series-resonant frequency of crystal "C" and with the crystals described, based on a 450-kc. carrier, will be approximately 448.6 kc.

5) Align all transformer trimmers for maximum response on this frequency.

6) Next, adjust the signal generator *slowly* in the higher-frequency direction until a *null* is obtained. This will be the series-resonant frequency of crystal "D," 450 kc. with the crystals indicated.

7) Move the signal generator ½ kc. *lower* than this null and adjust the trimmer on the input side of T_2 for maximum response.

8) Return signal generator to null.

9) Move the signal generator approximately 1 to 1.2 kc. *higher* than the null and adjust C_3 for *minimum* response.

10) Move the signal generator *higher* until another null is found: this will be the series-resonant frequency of crystal "B," approximately 452.8 kc. with the crystals shown.

11) Continue approximately ½ kc. higher than this null and adjust the output trimmer on T_1 slightly for moderate null.

12) Repeat Steps 7 through 11 to compensate for interaction, and alignment is complete.

For alignment of the dual filter the procedure is identical but must be done once for each sideband. However, when adjusting the filter for rejecting the *lower* sideband and where Steps 1–12 mention "higher" you must insert "lower" and vice versa. The alignment chart, Fig. 12-14, will simplify the alignment procedure on either filter.

The slug-tuned i.f. transformer is peaked at 3930 kc. and then stagger-tuned slightly to provide coverage of the entire 'phone band. The 6AG7 plate tank capacitor is adjustable from the front panel and is touched up when shifting frequency, as in the case of any transmitter amplifier stage.

A Two-Stage Linear Amplifier

The amplifier shown in Figs. 12-15, 12-17 and 12-18 is designed to follow a low-powered SSB exciter. As can be seen from the wiring diagram, Fig. 12-16, an 807 Class-A driver is used to excite a pair of 811-As operating Class B. Only a few watts is required to drive the 807, since it is never operated with grid current and the driving power is necessary only to overcome circuit losses. The 811-As will deliver about 180 watts peak with 1000 volts on the plates and 250 watts peak at 1200 volts. Operation as a linear amplifier for SSB with 1500 volts on the plates is not recommended because the driver stage is likely to introduce too much distortion, although a small amount of fixed bias (3–4½ volts) on the grids of the 811-As will permit c.w. operation at this higher plate voltage.

The circuit is not unlike ordinary Class-C practice, except for the bias voltages involved. The 807 stage uses cathode bias, and the 811-As run with zero bias (bias terminals short-circuited by a jumper wire). The most important factor in linear operation is the loading of the amplifiers, and thus provision has been made for varying the coupling on the 807 plate and the plates of the 811-As. The 807 loading is adjusted by varying the position of the link coil in L_3, and the link to L_6 is controlled from the front panel.

A low-inductance bypass condenser, C_2, made from a piece of coaxial line, helps to eliminate parasitics in the 807 stage, as does returning the screen bypass condenser, C_3, to the cathode instead of to ground. Grid chokes, L_4 and L_5, were found necessary to avoid high-frequency parasitic oscillations in the 811-A stage, as were resistors R_3, R_4 and R_5. All wiring other than r.f. was run in shield braid. Filament bypass condensers in the 811-A stage were found to be unnecessary.

Construction

The amplifier is built on a 13 by 17 by 3-inch aluminum chassis. The panel is an aluminum relay-rack panel, 15¾ inches high, that is held to

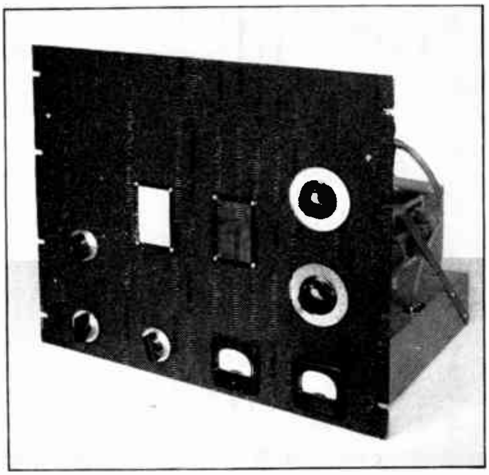


Fig. 12-15 — A two-stage linear amplifier for boosting the power level of a SSB signal. Large knobs control the antenna coupling and output plate tuning. The meters indicate grid and plate currents of the push-pull 811-As output stage.

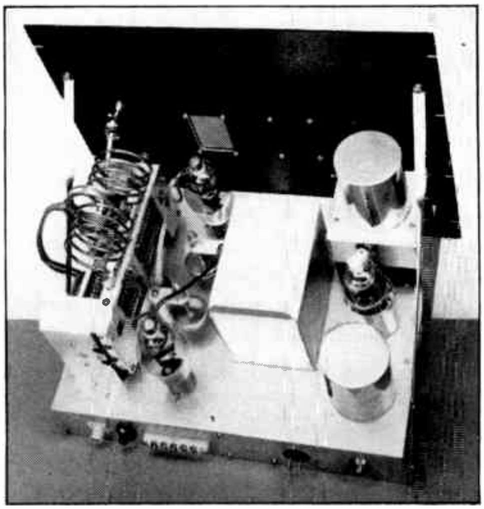


Fig. 12-17 — A rear view of the linear amplifier, showing the push-pull 811-A output amplifier and the 807 driver. The cover of the rectangular shield can slides off for access to the final grid coil. The round shield cans are for the 807 grid and plate coils.

the chassis by the shaft bearings and meters, and it is further braced by two strips of 1/16 by 1/2-inch brass.

The grid coil for the 807 plugs in to a socket mounted at the rear of the chassis and shielded by an ICA No. 1549 3-inch diameter shield can. The plate coil plugs in to a socket mounted 4 inches above the chassis. The platform for the socket also shields the plate condenser, C_5. Another 3-inch diameter shield can protects the 807 plate coil. The plate bypass condenser, C_6, is mounted under the chassis near the 807 socket, and the lead from C_5 and L_2 is brought down to it in shielded wire.

The grid coil for the 811-As is shielded by an ICA No. 29842 4 by 5 by 6 aluminum utility

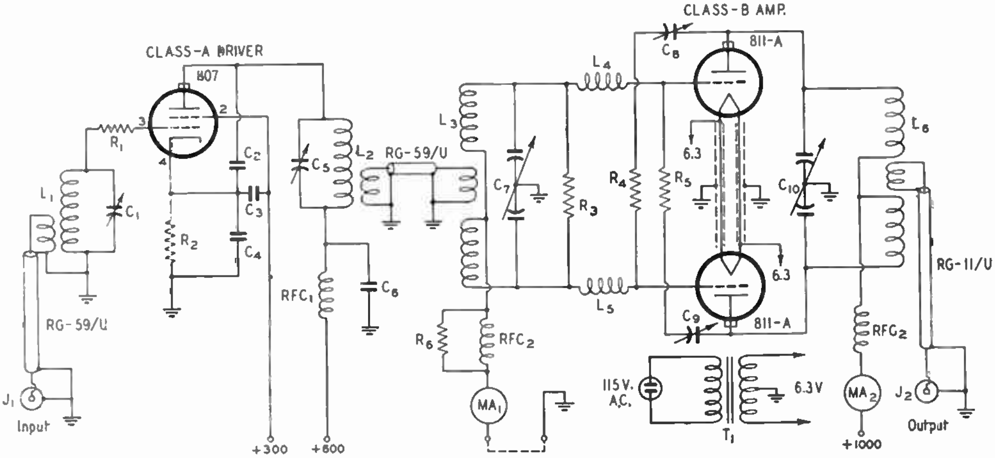


Fig. 12-16 — Wiring diagram of the linear amplifier.

C_1 — 140-μμfd. variable (Millen 19140).
C_2 — 13-μμfd. tubular, made of RG-58/U. Active length, 6 inches.
C_3, C_4 — .005-μfd. disc ceramic.
C_5 — 140-μμfd. variable (Millen 22140).
C_6 — .001-μfd. 1200-volt mica.
C_7 — Dual variable, 100-μμfd. per section (Millen 24100).
C_8, C_9 — Disc-type neutralizing condensers with feed-through base (Bud NC-853).
C_{10} — Dual variable, 200-μμfd. per section, .077-inch spacing (National MC-200D).
R_1 — 100 ohms, ½ watt.
R_2 — 680 ohms, 2 watts.
R_3 — 2700 ohms, 4 watts (4 2700-ohm in series-parallel).
R_4, R_5 — 20 ohms, 2 watts.
R_6 — 1000 ohms, 1 watt.
All resistors are composition, not wirewound.
J_1 — Input connector (Jones S-101-D).
J_2 — Coaxial-line connector (Amphenol 83-1R).
MA_1 — 0-50 milliammeter.
MA_2 — 0-500 milliammeter.
RFC_1 — 2.5-mh. 300-ma. r.f. choke.
RFC_2 — 200-μh. 75-ma. r.f. choke.
RFC_3 — 5-mh. 300-ma. r.f. choke (National R300S).
T_1 — 6.3-volt 100-amp. transformer (Stancor P-6308).

COIL TABLE FOR TWO-STAGE LINEAR AMPLIFIER

Band	*Turns*	*Wire No.*	*Diam.*	*Length*	*μh.*	*Link*	*Spacing*
L_1*							
3.9	22½	20 enam.	1	¾	10	4	1/16
14	10½	20 enam.	1	¾	2.5	3	1/16
L_2**							
3.9	25	20 enam.	1	⅞	11.2	4	1/16
14	11	20 enam.	1	¾	2.5	3	⅛
L_3***							
3.9	22	22 enam.	1¼	1¼	9.1	6	Adjustable
14	12	18 enam.	1¼	1⅛	3.3	4	Adjustable
L_6****							
3.9	22	16 enam.	2½	2¼	20	3	Adjustable
14	8	.15 tubing	2½	3¾	2.3	3	Adjustable

* Wound on Millen 45004 plug-in form.

** Wound on Millen 45005 plug-in form.

*** National AR-16-40S and AR-16-20S. 75-meter coil shunted by 150-μμfd. mica condenser.

**** B & W 80TVL with 18 turns removed, and B & W 15TVL.

cabinet. To simplify coil changing, the cabinet is fastened to the chassis and a friction-fit cover is made from a piece of sheet aluminum. The inside lips on the top of the cabinet should be bent down to allow more room for the hand that changes coils.

The output tank condenser, C_{10}, is mounted on the chassis with aluminum brackets that also support the jack bar for the output coil, L_6.

Adjustment

With a signal from the exciter coupled through J_1, and plate and screen voltages on the 807, it should be quite possible to drive the 811-A grid current off scale (with no plate voltage on the 811-As). Back off the excitation to about 25 ma. grid current and neutralize the 811-A stage by adjusting C_8 and C_9. The "flick" in grid current as C_{10} is tuned through resonance can be used, but a more sensitive indication, such as a crystal diode and 0–1 milliammeter connected to J_2, is to be preferred.

Couple a dummy load to J_2 and apply plate voltage to the 811-As. Couple an oscilloscope to the dummy load and apply a "two-tone" test signal to the unit, as described earlier in this chapter. The 811-A no-signal plate current should run around 40 or 50 ma., depending upon the plate voltage. Adjust the two-tone signal amplitude for 10 or 15 ma. grid current and resonate all circuits. Then increase the excitation until the two-tone pattern just begins to flatten on the peaks. When using 1000 volts on the 811-As, this flattening should not occur before MA_2 indicates 160 ma. or so — with 1200 volts the current should run up to 190 ma. without noticeable flattening. If flattening occurs sooner, it indicates that the 811-A stage should be coupled more tightly to its load, or that the 807 stage is not delivering enough drive. It will probably be found that the 811-A output coupling is at fault — if the link is coupled closely at L_3 the 807 should behave at all power levels.

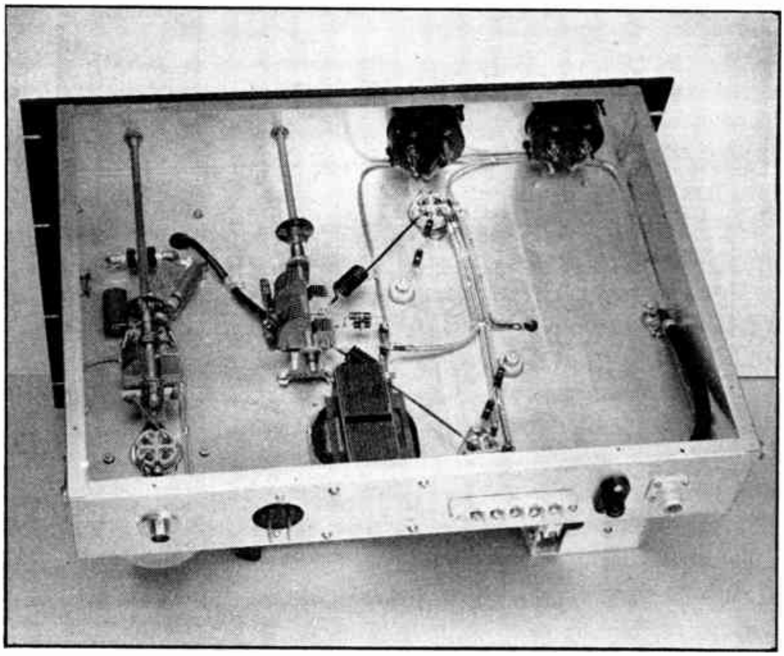


Fig. 12-18 — Underneath the chassis, showing all but r.f. leads in shield braid. The coils in the leads from the split-stator grid condenser are parasitic chokes.

CHAPTER 13

Transmission Lines

The place where r.f. power is generated is very frequently not the place where it is to be utilized. A transmitter and its antenna are a good example: The antenna, to radiate well, should be high above the ground and should be kept clear of trees, buildings and other objects that might absorb energy, but the transmitter itself is most conveniently installed indoors where it is readily accessible. There are numerous other instances where power must be delivered from one point to another, even though the distance may be only a few feet.

The means by which power is transported from one spot to another is the r.f. transmission line. At radio frequencies a line exhibits entirely different characteristics than it does at commercial power frequencies. This is because the speed at which electrical energy travels, while tremendously high as compared with mechanical motion, is not infinite. The peculiarities of r.f. transmission lines result from the fact that an interval comparable with the time of an r.f. cycle must elapse before energy leaving one point in the circuit can reach another just a short distance away.

The discussion to follow assumes that the line consists of two parallel wires, separated by a distance very small compared with the wavelength. The parallel-conductor line is not the only type, but the same principles apply to all varieties of lines.

Operating Principles

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 each wire 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. 13-1, 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.

Characteristic Impedance

An infinitely-long chain of coils and condensers connected as in Fig. 13-1, where each L is the same as all others and all the C's have the same value, has an important property. 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 $\sqrt{L/C}$, where L and C are the inductance and capacitance per unit length. This impedance is purely resistive.

In defining the characteristic impedance as $\sqrt{L/C}$, it is assumed that the conductors have no inherent resistance — that is, there is no I^2R loss in them — and that there is no power loss in the dielectric surrounding the conductors. In other words, it is assumed there is no power loss in or from the line no matter how great its length. This does not seem consistent with calling the characteristic impedance a pure resistance, which implies that power supplied is all dissipated in the line. But in an infinitely-long line the effect, so far as the source of power is concerned, is exactly the same as though the power were dissipated in a resistance, because the power leaves the source and travels outward forever along the line.

The characteristic impedance determines the amount of current that can flow when a given voltage is applied to an infinitely-long

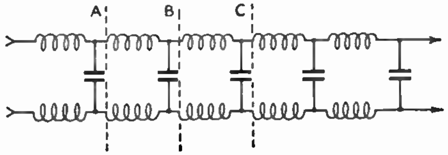


Fig. 13-1 — **Equivalent of a transmission line in lumped circuit constants.**

line, in exactly the same way that a definite value of actual resistance limits current flow when a given voltage is applied.

The inductance and capacitance per unit length of line depend upon the size of the conductors and the spacing between them. The closer the two conductors 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.

"Matched" Lines

Actual transmission lines do not extend to infinity but have a definite length and are connected to, or **terminate** 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 with 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 energy will travel a distance of one wavelength along the line wires. 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. Hence 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. However, at any given point along the line the current goes through similar variations with time 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 of travel 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

In the infinitely-long line (or its matched counterpart) the impedance is the same at any point on the line because the ratio of voltage to current is always the same. However, the impedance at the end of the line in Fig. 13-2 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, reverses its direction of flow and goes back along the transmission line toward the input end. There is a 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.

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 will be doubled. At in-between points the amplitude is between these two extremes. The points

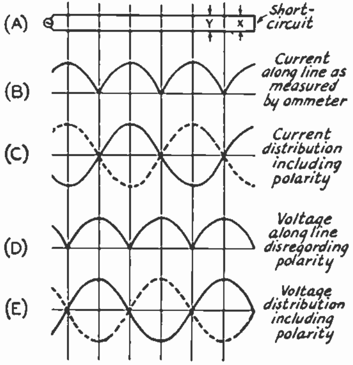


Fig. 13-2 — Standing waves of voltage and current along a short-circuited transmission line.

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.

In the short-circuit at the end of the line the two current components are in phase and the total current is large. 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 resultant current will again 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 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. 13-2B. 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. 13-2C. Furthermore, the current in the second wire is flowing in the opposite direction to the current in the adjacent section of the first wire. This is indicated by the broken curve in Fig. 13-2C. 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** or **current antinode** 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 a reflected voltage of equal amplitude and 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. 13-2, 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 is called a **voltage loop** or **antinode** and a voltage minimum is called a **voltage node.**

Open-Circuited Line

If the end of the line is open-circuited instead of short-circuited, there can be no current at the end of the line but a large voltage can exist. Again the outgoing power is reflected back toward the source. 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 and add together. The result is that we again have standing waves, but the conditions are reversed as compared with a short-circuited line. Fig. 13-3 shows the open-circuited line case.

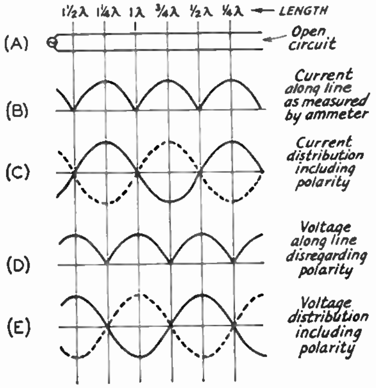


Fig. 13-3 — Standing waves of current and voltage along an open-circuited transmission line.

Lines Terminated in Resistive Load

Fig. 13-4 shows a line terminated in a resistive load. In this 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 neither voltage nor current cancel completely 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 resistance, Z_r, is equal to the characteristic impedance, Z_0, of the line all the power is absorbed in the load. In such a case there is no reflected power and therefore no standing waves of current and voltage. This is a special case that represents the change-over point between "short-circuited" and "open-circuited" lines. If Z_r is less than Z_0, the current is largest at the load, while if Z_r is greater than Z_0 the voltage is largest at the load. The two conditions are shown at B and C, respectively, in Fig. 13-4.

The resistive termination is an important practical case. The termination is seldom an

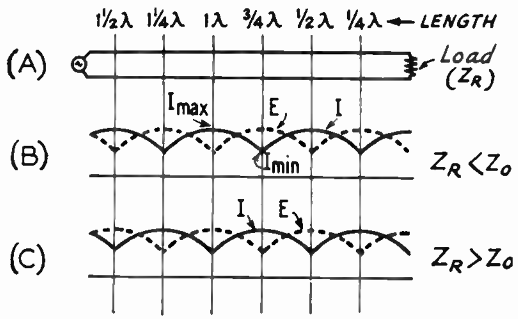


Fig. 13-4 — Standing waves on a transmission line terminated in a resistive load.

actual resistor, the most common terminations being resonant circuits or resonant antenna systems, both of which have essentially resistive impedances. If the load is reactive as well as resistive, the operation of the line resembles that shown in Fig. 13-4, but the presence of reactance in the load causes two modifications: The loops and nulls are shifted toward or away from the load; and the amount of power reflected back toward the source is increased, as compared with the amount reflected by a purely resistive load of the same total impedance. Both effects become more pronounced as the ratio of reactance to resistance in the load is made larger.

Standing-Wave Ratio

The ratio of maximum current to minimum current along a line, Fig. 13-5, is called the **standing-wave ratio.** The same ratio holds for maximum voltage and minimum voltage. It is a measure of the mismatch between the load and the line, and is equal to 1 when the line is perfectly matched. (In that case the "maximum" and "minimum" are the same, since the current and voltage do not vary along the line.) When the line is terminated in a purely-resistive load, the standing-wave ratio is

$$S.W.R. = \frac{Z_r}{Z_0} \text{ or } \frac{Z_0}{Z_r} \qquad \textbf{(13-A)}$$

Where $S.W.R.$ = Standing-wave ratio

Z_r = Impedance of load (must be pure resistance)

Z_0 = Characteristic impedance of line

Example: A line having a characteristic impedance of 300 ohms is terminated in a resistive load of 25 ohms. The s.w.r. is

$$S.W.R. = \frac{Z_0}{Z_r} = \frac{300}{25} = 12 \text{ to } 1$$

It is customary to put the larger of the two quantities, Z_r or Z_0, in the numerator of the fraction so that the s.w.r. will be expressed by a number larger than 1.

It is easier to measure the standing-wave ratio than some of the other quantities (such as the impedance of an antenna) that enter into transmission-line computations. Consequently, the s.w.r. is a convenient basis for work with lines. The higher the s.w.r., the greater the mismatch between line and load. In practical lines, the power loss in the line itself increases with the s.w.r.

● INPUT IMPEDANCE

The input impedance of a transmission line is the impedance seen looking into the sending-end or input terminals; it is the impedance into which the source of power must work when the line is connected. If the load is perfectly matched to the line the line appears to be infinitely long, as stated earlier, and the input impedance is simply the characteristic impedance of the line itself. However, if there are standing waves this is no longer true; the input impedance may have a wide range of values.

This can be understood by referring to Figs. 13-2, 13-3, or 13-4. If the line length is such that standing waves cause the voltage at the input terminals to be high and the current low, then the input impedance is higher than the Z_0 of the line, since impedance is simply the ratio of voltage to current. Conversely, low voltage and high current at the input terminals mean that the input impedance is lower than the line Z_0. Comparison of the three drawings also shows that the range of input impedance values that may be encountered is greater when the far end of the line is open- or short-circuited than it is when the line has a resistive load. In other words, the higher the s.w.r. the greater the range of input impedance values when the line length is varied.

In addition to the variation in the absolute value of the input impedance with line length, the presence of standing waves also causes the input impedance to contain both reactance and resistance, even though the load itself may be a pure resistance. The only exceptions to this occur at the exact current loops or nodes, at which points the input impedance is a pure resistance. These are the only points at which the outgoing and reflected voltages and currents are exactly in phase: At all other distances along the line the current either leads or lags behind the voltage and the effect is exactly the same as though a capacitance or

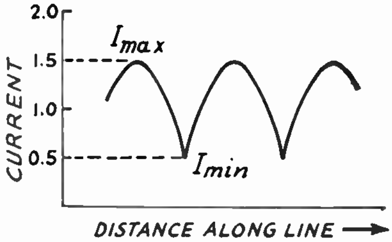


Fig. 13-5 — Measurement of standing-wave ratio. In this drawing, I_{max} is 1.5 and I_{min} is 0.5, so the s.w.r. = I_{max}/I_{min} = 1.5/0.5 = 3 to 1.

inductance were part of the input impedance of the line.

The input impedance can be represented by either a resistance and a capacitance, or as a resistance and an inductance, as shown in Fig. 13-6. Whether the impedance is inductive or capacitive depends on the characteristics of the load and the length of the line. It is possible to represent the equivalent circuit by resistance and reactance either in series or parallel, so long as the total impedance and phase angle are the same in either case. Meeting this last condition requires different values of resistance and reactance in the series case than in the parallel case.

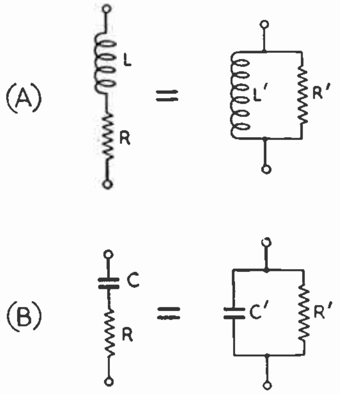


Fig. 13-6 — Series and parallel equivalents of a line whose input impedance has both reactive and resistive components. The series and parallel equivalents do not have the same values; e.g., in A, *L* does not equal *L'* and *R* does not equal *R'*.

The magnitude and character of the input impedance is quite important, since it determines the method by which the power source must be coupled to the line. The calculation of input impedance is rather complicated and its measurement is not feasible with ordinary equipment. Fortunately, in amateur work, it is unnecessary either to calculate or measure it. The proper coupling can be achieved by relatively simple methods described later in this chapter.

Unterminated Lines

The input impedance of a short-circuited or open-circuited line not an exact multiple of one-quarter wavelength long is practically a pure reactance. This is because there is very little power lost in the line. Such lines are frequently used as "linear" inductances and capacitances.

If a shorted line is less than a quarter wave long, as at *X* in Fig. 13-2, it will have inductive reactance. The reactance increases with the line length up to the quarter-wave point. Beyond that, as at *Y*, the reactance is capacitive, high near the quarter-wave point and becoming lower as the half-wave point is approached. It then alternates between inductive and capacitive in successive quarter-wave sections. Just the reverse is true of the open-circuited line.

At exact multiples of a quarter wavelength the impedance is purely resistive. It is apparent, from examination of B and D in Fig. 13-2, 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.)

Impedance Transformation

The fact that the input impedance of a line depends on the s.w.r. and line length can be used to advantage when it is necessary to transform a given impedance into another value.

Study of Fig. 13-4 will show that, just as in the open- and short-circuited cases, if the line is one-half wavelength long the voltage and current are exactly the same at the input terminals as they are at the load. This is also true of lengths that are integral multiples of a half wavelength. It is also true for all values of s.w.r. Hence the input impedance of any line, no matter what its Z_0, that is a multiple of a half-wavelength long is exactly the same as the load impedance. Such a line can be used to transfer the impedance to a new location without changing its value.

When the line is a quarter wavelength long, or an odd multiple of a quarter wavelength, the load impedance is "inverted." That is, if the current is low and the voltage is high at the load, the input impedance will be such as to require high current and low voltage. The relationship between the load impedance and input impedance is given by:

$$Z_s = \frac{Z_0^2}{Z_r} \qquad \text{(13-B)}$$

where Z_s = Impedance looking into line (line length an odd multiple of one-quarter wavelength)

Z_r = Impedance of load (must be pure resistance)

Z_0 = Characteristic impedance of line

Example: A quarter-wavelength line having a characteristic impedance of 500 ohms is terminated in a resistive load of 75 ohms. The impedance looking into the input or sending end of the line is

$$Z_s = \frac{Z_0^2}{Z_r} = \frac{(500)^2}{75} = \frac{250{,}000}{75} = 3333 \text{ ohms}$$

If the formula above is rearranged, we have

$$Z_0 = \sqrt{Z_s Z_r} \qquad (13\text{-C})$$

This means that if we have two values of impedance that we wish to "match," we can do so if we connect them together by a quarter-wave transmission line having a characteristic impedance equal to the square root of their product. A quarter-wave line, in other words, has the characteristics of a transformer.

Resonant and Nonresonant Lines

Because the input impedance of a line operating with a high s.w.r. is critically dependent on the line length, and furthermore is usually reactive as well as resistive, special tuning means are required for effective power transfer from the source to the line. Lines operated in this way are commonly called "tuned" or "resonant" lines. On the other hand, if the s.w.r. is low the input impedance is close to the Z_0 of the line and does not vary a great deal with the line length. Such lines are called "flat," or "untuned", or "nonresonant".

There is no sharp line of demarkation between tuned and untuned lines. If the s.w.r. is below 1.5 to 1 the line is essentially flat, since the same coupling method will work with all line lengths. If the s.w.r. is above 3 or 4 to 1 the type of coupling system, and its adjustment, will depend on the line length and such lines fall into the "tuned" category.

It is always advantageous to make the s.w.r. as low as possible. "Tuning the line" becomes necessary only when a considerable mismatch between the load and the line has to be tolerated. The most important practical example of this is when a single antenna is operated on several harmonically-related frequencies, in which case the antenna impedance will have widely-different values on different harmonics.

● RADIATION

Whenever a wire carries alternating current the electromagnetic fields travel away into space with the velocity of light. At power-line frequencies the field that "grows" when the current is increasing has plenty of time to return or "collapse" about the conductor when the current is decreasing, because the alternations are so slow. But at radio frequencies fields that travel only a relatively short distance do not have time to get back to the conductor before the next cycle commences. The consequence is that some of the electromagnetic energy is prevented from being restored to the conductor; in other words, energy is radiated into space in the form of electromagnetic waves.

The amount of energy radiated depends, among other things, on the length of the conductor in relation to the frequency or wavelength of the r.f. current. If the conductor is very short compared to the wavelength the energy radiated will be small. However, a transmission line used to feed power to an antenna is not short in this sense; in fact, it is almost always an appreciable fraction of a wavelength long and may have a length of several wavelengths.

The lines previously considered have consisted of two parallel conductors of the same diameter. Provided there is nothing in the system to destroy symmetry, at every point along the line the current in one conductor has the same intensity as the current in the other conductor at that point, but the currents flow in opposite directions. This was shown in Figs. 13-2C and 13-3C. It means that the fields set up about the two wires have the same intensity, but *opposite directions*. The consequence is that the total field set up about such a transmission line is zero; the two fields "cancel out." Hence no energy is radiated.

Actually, the fields do not completely cancel out because for them to do so the two conductors would have to occupy the same space, whereas they are slightly separated. However, the cancellation is substantially complete if the distance between the conductors is very small compared to the wavelength. Radiation will be negligible if the distance between the conductors is 0.01 wavelength or less, provided the currents in the two actually are balanced as described.

The amount of radiation also is proportional to the current flowing in the line. Because of the way in which the current varies along the line when there are standing waves, the effective current, for purposes of radiation, becomes greater as the s.w.r. is increased. For this reason the radiation is least when the line is flat. However, if the conductor spacing is small and the currents are balanced, the radiation from a line with even a high s.w.r. is inconsequential. A small unbalance in the line currents is far more serious.

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 the construction and use of parallel and coaxial lines.

● PARALLEL-CONDUCTOR LINES

A common type of parallel-conductor line used in amateur installations is one 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. 13-7. Such a line is said to be **air-insulated.** Typical spacers are shown in Fig. 13-8. 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.

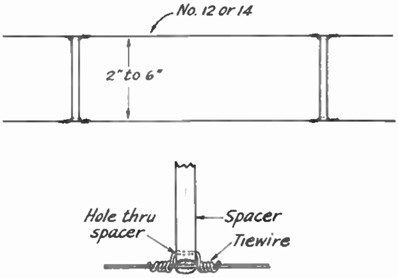


Fig. 13-7 — Typical construction of open-wire line. The line conductor fits in a groove in the end of the spacer, and is held in place by a tie-wire anchored in a hole near the groove.

Prefabricated parallel-conductor line with air insulation has been developed as a low-loss line for television reception and can also be used in transmitting applications. This line consists of two No. 18 conductors held at a spacing of one inch by molded-on spacers. The characteristic impedance is 450 ohms.

A convenient type of manufactured line is one in which the parallel conductors are imbedded in low-loss insulating material (polyethylene). It is commonly used as a TV lead-in and has a characteristic impedance of 300 ohms. It is sold under various names, the most common of which is "Twin-Lead". This type of line has the advantages of light weight, close and uniform conductor spacing, flexibility and neat appearance. However, the losses in the solid dielectric are higher than in air, and dirt or moisture on the line tends to change the characteristic impedance. Moisture effects can be reduced by coating the line with silicone grease. A special form of 300-ohm Twin-Lead for transmitting uses a polyethylene tube with the conductors molded diametrically opposite; the longer dielectric path in such line reduces moisture troubles.

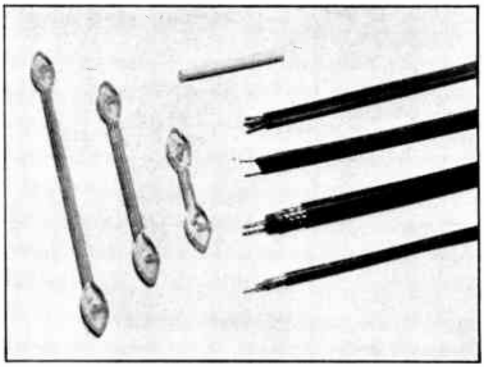


Fig. 13-8 — Typical manufactured transmission lines and spacers.

In addition to 300-ohm line, Twin-Lead is obtainable with a characteristic impedance of 75 ohms for transmitting purposes. Lightweight 75- and 150-ohm Twin-Lead also is available.

Characteristic Impedance

The characteristic impedance of an air-insulated parallel-conductor line is given by:

$$Z_0 = 276 \log \frac{b}{a} \qquad \textbf{(13-D)}$$

where Z_0 = Characteristic impedance
b = Center-to-center distance between conductors
a = Radius of conductor (in same units as b)

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

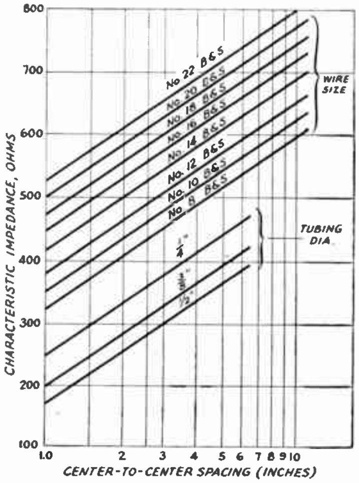


Fig. 13-9 — Chart showing the characteristic impedance of spaced-conductor parallel transmission lines with air dielectric. Tubing sizes given are for outside diameters.

in Fig. 13-9 for a number of common conductor sizes.

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.

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.

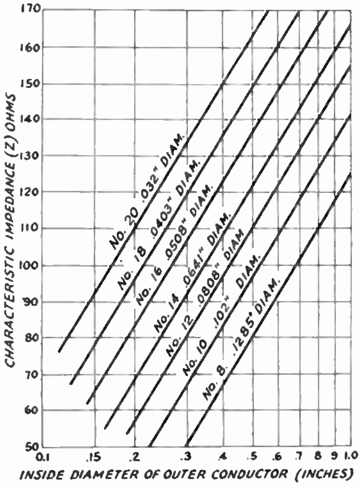


Fig. 13-10 — Chart showing characteristic impedance of various air-insulated concentric lines.

COAXIAL LINES

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. 13-8. 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 coaxial line is given by the formula

$$Z_0 = 138 \, log \frac{b}{a} \qquad \textbf{(13-E)}$$

where Z_0 = Characteristic impedance
b = Inside diameter of outer conductor
a = Outside diameter of inner conductor (in same units as b)

Curves for typical conductor sizes are given in Fig. 13-10.

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.

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 material dielectrics 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;

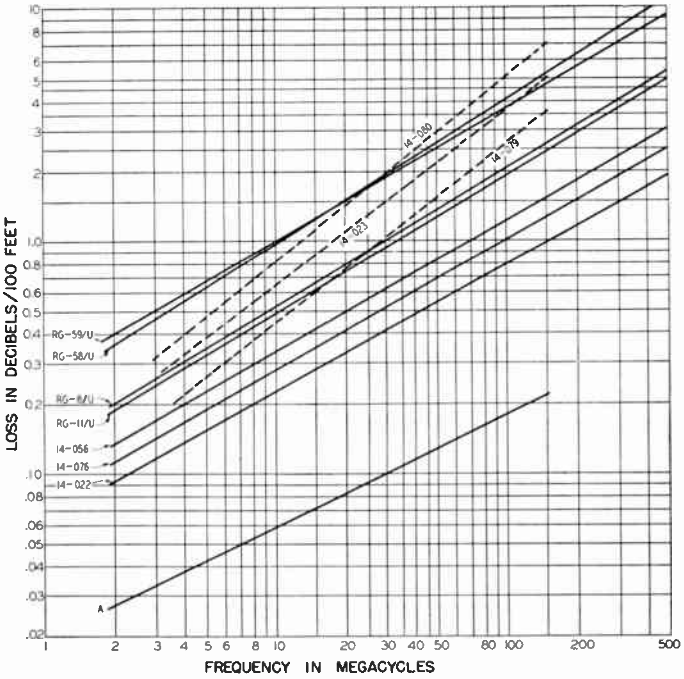


Fig. 13-11 — Attenuation data for common types of transmission lines. Curve A is the nominal attenuation of 600-ohm open-wire line with No. 12 conductors, not including dielectric loss in spacers nor possible radiation losses. Additional line data are given in Table 13-I.

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.

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

$$Length\ in\ feet = \frac{984}{f} \cdot V \qquad \textbf{(13-F)}$$

where f = Frequency in megacycles
V = Velocity factor

TABLE 13-I

Transmission-Line Data

Type	Description or Type Number	Characteristic Impedance	Velocity Factor	Capacitance per foot; μμfd.
Coaxial	Air-insulated	50–100	0.85[1]	
	RG-8/U	53	0.66	29.5
	RG-58/U	53	0.66	28.5
	RG-11/U	75	0.66	20.5
	RG-59/U	73	0.66	21.0
Parallel-Conductor	Air-insulated	200–600	0.975[2]	
	14–080[3]	75	0.68	19.0
	14–023[3]	75	0.71	20.0
	14–079[3]	150	0.77	10.0
	14–056[3]	300	0.82	5.8
	14–076[3]	300	0.84	3.9
	14–022[3]	300	0.85	3.0

[1] Average figure for small-diameter lines with ceramic beads.

[2] Average figure for lines insulated with ceramic spacers at intervals of a few feet.

[3] Amphenol type numbers and data. Line similar to 14–056 is made by several manufacturers, but rated loss may differ from that given in Fig. 13-11. Types 14–023, 14–076, and 14–022 are made for transmitting applications.

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 13-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 13-I, V is 0.82. At this frequency (7.15 Mc.) a wavelength is

$$Length\ (\text{feet}) = \frac{984}{f} \cdot V = \frac{984}{7.15} \times 0.82$$
$$= 137.6 \times 0.82 = 112.8\ \text{ft.}$$

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

$$Length\ (\text{feet}) = \frac{246}{f} \cdot V \qquad \textbf{(13-G)}$$

where the symbols have the same meaning as above.

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. There is no appreciable radiation loss from a coaxial line, but radiation from a parallel-conductor line may exceed the heat losses if the line is un-

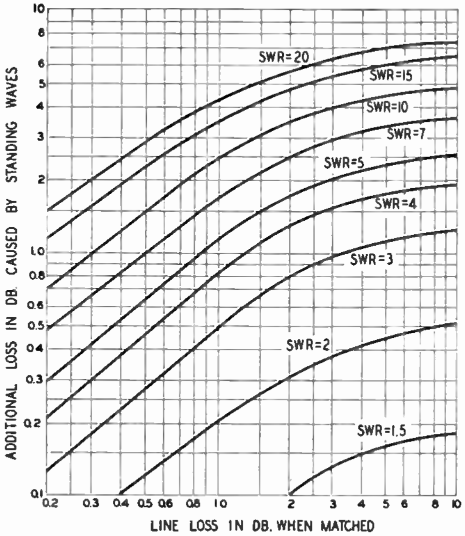


Fig. 13-12 — **Effect of standing-wave ration on line loss. The ordinates give the *additional* loss in decibels for the loss, under perfectly-matched conditions, shown on the horizontal scale.**

balanced. Since radiation losses cannot readily be estimated or measured, the following discussion is based only on conductor and dielectric losses.

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.

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 graphical form in Fig. 13-11. In these curves the radiation loss is assumed to be negligible.

When there are standing waves on the line the power loss increases as shown in Fig. 13-12. 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 Fig. 13-11 would be 1.5 × 0.4 = 0.6 db. From Fig. 13-12 the additional loss because of the s.w.r. is 0.73 db. The total loss is therefore 0.6 + 0.73 = 1.33 db.

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. 13-12, 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.

Matching the Load to the Line

The load for a transmission line may be any device capable of dissipating r.f. power. When lines are used for transmitting applications the most common type of load is an antenna, but there are also practical cases where the grid circuit of a power amplifier may represent the load. When a transmission line is connected between an antenna and a receiver, the receiver input circuit (not the antenna) is the load, because the power taken from a passing wave is delivered to the receiver.

Whatever the application, the conditions existing at the load, and only the load, determine the standing-wave ratio on the line. If the load is purely resistive and equal in value to the characteristic impedance of the line, there will be no standing waves. If the load is not purely resistive, and/or is not equal to the line Z_0, there will be standing waves. No adjustments that can be made at the input end of the line can change the s.w.r., nor is it affected by changing the line length.

Only in a few special cases is the load inherently of the proper value to match a practicable transmission line. In all other cases it is necessary either to operate with a mismatch and accept the s.w.r. that results, or else to take steps to bring about a proper match between the line and load by means of transformers or similar devices. Impedance-matching transformers may take a variety of physical forms, depending on the circumstances.

Note that it is essential, if the s.w.r. is to be made as low as possible, that the load at the point of connection to the transmission line be purely resistive. In general, this requires that the load be tuned to resonance. If the load itself is not resonant at the operating frequency the tuning sometimes can be accomplished in the matching system.

● THE ANTENNA AS A LOAD

Every antenna system, no matter what its physical form, will have a definite value of impedance at the point where the line is to be connected. The problem is to transform this "antenna input impedance" to the proper value to match the line. In this respect there is no one "best" type of line for a particular antenna system, because it is possible to transform impedances in any desired ratio. Consequently, any type of line may be used with any type of antenna. (There are frequently reasons other than impedance matching that dictate the use of one type of line in preference to another, such as ease of installation, inherent loss in the line, and so on, but these are not considered in this section.)

Although the input impedance of an antenna system is seldom known very accurately, it is often possible to make a reasonably close estimate of its value. The information in the chapter on antennas can be used as a guide.

Matching circuits may be constructed using ordinary coils and condensers, but are not used very extensively because they must be supported at the antenna and must be weatherproofed. The systems to be described use "linear" transformers.

The Quarter-Wave Transformer or "Q" Section

As described earlier in this chapter, a quarter-wave transmission line may be used as an impedance transformer. Knowing the antenna impedance and the characteristic impedance of the transmission line to be matched, the required characteristic impedance of a matching section such as is shown in Fig. 13-13 is

$$Z_m = \sqrt{Z_1 Z_2}$$

where Z_1 is the antenna impedance and Z_2 is the characteristic impedance of the line to which it is to be matched.

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. 13-9. (With ½-inch tubing, the spacing in the example above should be 1.5 inches for an impedance of 208 ohms.)

The length of the quarter-wave matching section is given by Equation 13-G.

The antenna must be resonant at the operating frequency. Setting the antenna length by formula is amply accurate with single-wire antennas, but in other systems, particularly close-spaced arrays, the antenna should be adjusted to resonance before the matching section is connected.

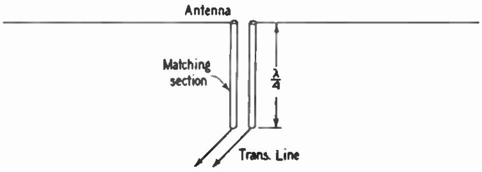


Fig. 13-13 — "Q" matching section, a quarter-wave impedance-transformer.

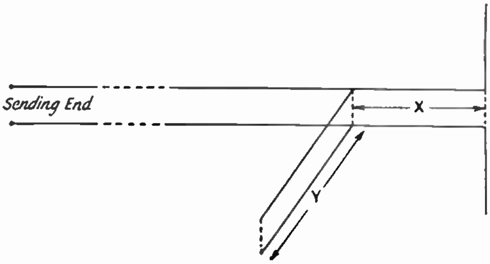


Fig. 13-14 — **Matching the antenna to the line by means of a stub, Y. Curves for determining the lengths X and Y are given in Figs. 13-15 and 13-16, for the case where the line, section X and section Y all have the same characteristic impedance.**

When the antenna input impedance is not known accurately, it is advisable to construct the matching section so that the spacing between conductors can be changed. The spacing then may be adjusted to give the lowest s.w.r. on the transmission line.

Stub Matching

When a transmission line is not matched by the load, the impedance looking into the line toward the load varies with the distance from the load, as discussed earlier in this chapter. Considering the input impedance to be equivalent to a resistance in parallel with a reactance, at some distance along the line, such as X in Fig. 13-14, the resistive part of the input impedance will be equal to the Z_0 of the line. If at this point a reactance equal to the reactive part of the input impedance, but of the opposite type, is connected across the line, the reactances will cancel and leave only the resistive component. From this point back to the transmitter or other source of energy the line will be matched.

The reactances used for matching in this way are usually "linear" reactances — sections of transmission line — called "stubs". Stubs may be "open" or "closed," depending on whether the free end is left open or is short-circuited, according to the type of reactance required in a particular case. The type and length of stub, as well as the point at which it should be attached to the line, can be found without any knowledge of the antenna input impedance, providing that the s.w.r. on the line can be measured before the stub is attached, and providing that the position of a current node (voltage loop) can be determined under the same conditions.

When the s.w.r. and the position of a current node are known Figs. 13-15 and 13-16 give the stub information necessary for impedance matching. Stub lengths are given in wavelengths, which may be converted to feet with the help of Equation 13-F. The data in Figs. 13-15 and 13-16 are based on the assumption

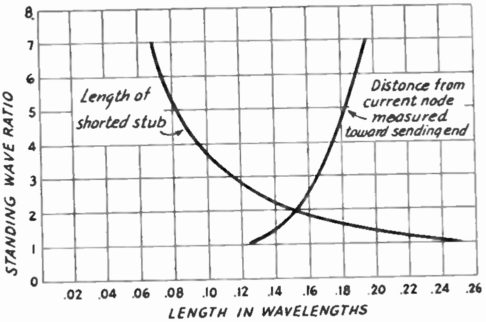


Fig. 13-15 — Graph for determining position and length of a *shorted* stub. Dimensions may be converted to linear units after values have been taken from the graph.

that the line and stub both have the same Z_0.

With this system of matching it is not necessary that the antenna system be exactly resonant, since the match is based on the position of a current node along the line. The node nearest the antenna should be used for determining the position of the stub so that as much as possible of the transmission line will be operating with a low s.w.r.

Study of the curves in Figs. 13-15 and 13-16 will show that when the initial s.w.r. is high (over 4 to 1) the sum of the stub length and distance from a current node is very close to 0.25 wavelength in the case of the closed stub and to 0.5 wavelength in the case of the open stub. In such cases the system may be visualized as shown in Figs. 13-17, as though a quarter-wave section of line formed a transformer along which the main transmission line can be tapped for impedance matching. When using this concept the antenna system should first be resonated to the operating frequency without the matching section attached. The positions of the line taps on the matching section are then adjusted to give the lowest possible s.w.r. on the feed line.

Folded Dipoles

A half-wave antenna element itself may be used to match various line impedances if it is split into two or more parallel conductors with the transmission line attached at the center of only one of them. Various forms of such "folded dipoles" are shown in Fig. 13-18. Currents in all conductors are in phase in a folded dipole, and since the conductor spacing is small the folded dipole is equivalent in radiating properties to an ordinary single-conductor dipole. The current flowing into the input terminals of the antenna from the line is the current in one conductor only, but the entire power from the line is delivered at this value of current. This is equivalent to saying that the input impedance of the antenna has been raised by splitting it up into two or more conductors.

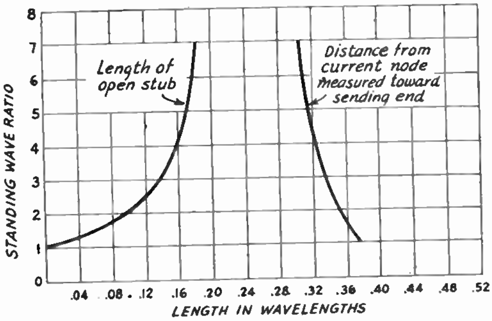


Fig. 13-16 — Graph for determining position and length of an *open* stub. Dimensions may be converted to linear units after values have been taken from the graph.

If the conductors of a folded dipole are all the same diameter and the spacing between them is small, the impedance at the input terminals is approximately equal to the input impedance of an ordinary dipole multiplied by the square of the number of conductors. A simple half-wave antenna has an average impedance of 70 ohms, so a two-conductor folded dipole will have an input impedance of 280 ohms, and a three-conductor dipole an impedance of 630 ohms. These values are sufficiently close for good matching to 300-ohm or 600-ohm line, respectively.

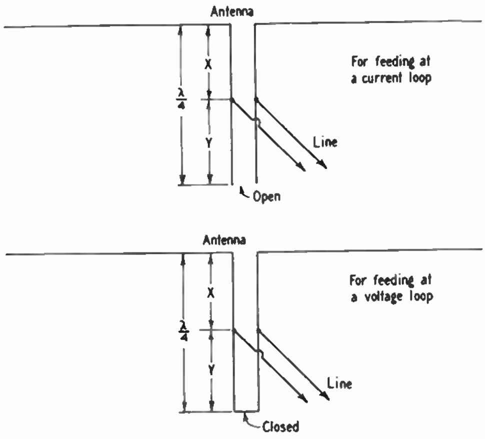


Fig. 13-17 — Matching by means of quarter-wave linear transformers.

Other values of impedance ratio may be obtained by making one conductor larger in diameter than the other, as shown at C in Fig. 13-18. The required ratio of conductor radii (or diameters) for a desired impedance ratio may be obtained from Fig. 13-19. The unequal-conductor method has been found particularly useful in matching to low-impedance antennas such as directive arrays using close-spaced parasitic elements.

The length of the antenna element should be such as to be approximately self-resonant at the median operating frequency. The length is usually not highly critical, because this method of matching tends to compensate for changes in antenna reactance with frequency

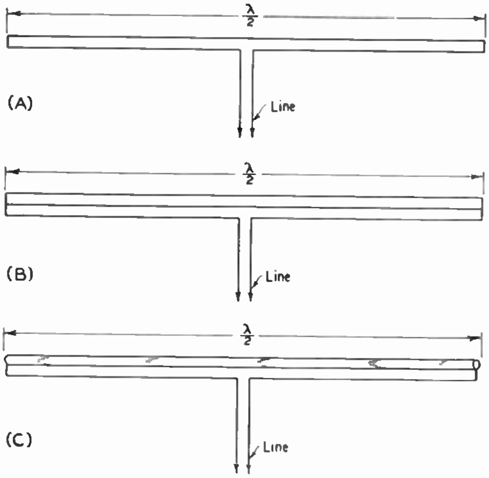


Fig. 13-18 — The folded dipole, a method for using the antenna element itself to provide an impedance transformation.

and thus broadens the frequency-response curve of the antenna.

"T" and "Gamma" Matching Sections

The method of matching shown in Fig. 13-20A is based on the fact that the impedance between any two points symmetrically located with respect to the center of a resonant antenna is resistive, and has a value which depends on the spacing between the two points. It is therefore possible to choose a pair of points between which the impedance will have the right value to match a transmission line. In practice, the line cannot be connected directly at these points because the distance between them is much greater than the conductor spacing of a practicable transmission line. The "T" arrangement in Fig. 13-20A overcomes this difficulty by using a second conductor paralleling the antenna to form a matching section to which the line may be connected.

The distance between the points of connection (twice the distance y in the drawing) does not represent the same impedance as the Z_0 of the transmission line, since the matching section introduces a degree of impedance transformation that depends on the length and diameter of the conductors and their distance from the antenna. Also, the matching section probably shows some reactance at its input terminals because it is not the proper length to be resonant. Design information that will take care of all these factors is not available, hence the system has to be adjusted by cut-and-try. The method of adjustment commonly used is to make the spacing x some value that is convenient constructionally, and then adjust the distance y, while maintaining symmetry with respect to the center, until the s.w.r. on the transmission line is as low as possible. If the s.w.r. is not below 2 to 1 after this adjustment, the antenna length should be changed slightly (lengthening is frequently necessary) and the matching-section taps adjusted again.

The unbalanced ("gamma") arrangement in Fig. 13-20B is similar in principle to the "T," but is adapted for use with single coax line. The method of adjustment is the same.

Dimensions of matching sections in practical cases are given in the chapter on antennas.

The "Delta" Match

The matching system in Fig. 13-21 is based on the variation in impedance between two points symmetrically located with respect to the center of the antenna, as in the case of the "T" match, but uses a different matching section. If the two conductors of a transmission line are fanned out, the Z_0 of the line will increase with the increase in spacing. A fanned section of line can be used to match a given load impedance to the Z_0 of a uniformly-spaced transmission line, provided the line Z_0 is lower than the impedance of the load.

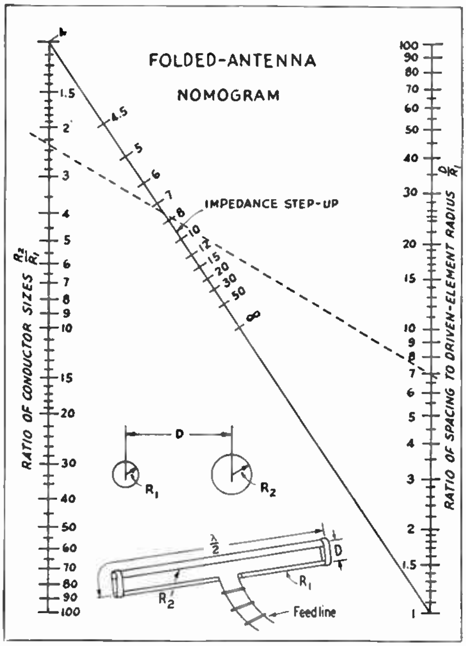


Fig. 13-19 — Nomogram for computing impedance step-up in a folded dipole with dissimilar conductors. The line at the left is the ratio of conductor diameters, and the line at the right is the ratio of conductor spacing (center-to-center) to the radius of the smaller (driven) conductor. The solid slanting line gives the impedance step-up ratio. Laying a straightedge between any two known quantities will give the value of the third.

Example: Find the spacing required when the diameter of the large conductor is 1.15 inch, the diameter of the smaller conductor is 0.5 inch, line impedance 300 ohms, antenna impedance 40 ohms.

Impedance step-up required = 300/40 = 7.5.
Ratio of conductor sizes = 1.15/0.5 = 2.3.
Laying a straightedge across the figure (dashed line), ratio of spacing to radius = 7.
Spacing, D = 7 × 0.25 = 1.75″.

Strictly, such a match can be made only if the conductor spacing in the fanned section of line increases at an exponential rate, but the "delta" arrangement in Fig. 13-21 approximates this type of spacing.

Dimensions *a* and *b* in Fig. 13-21 depend on the antenna impedance (whether it is a simple half-wave antenna or the driven element of a multielement beam), the size of the conductors in the delta, and the Z_0 of the transmission line to be matched. Methods for calculation are not available, but dimensions for practical cases are given in the chapters on antennas.

NONRADIATING LOADS

Important practical cases of nonradiating loads for a transmission line are the grid circuit of a power amplifier, the input circuit of a receiver, and another transmission line. This last case includes the "antenna tuner" (a misnomer because it is actually a device for coupling a transmission line to the transmitter) which, because of its importance in amateur installations, is considered separately in a later section of this chapter.

P. A. Grid Circuits

When link coupling is used between a driver and a power amplifier it is possible to look upon the link as a means for obtaining mutual inductance, as described in the chapters on electrical fundamentals and transmitters, if the link is electrically short. However, by treating the link as a transmission line it becomes possible to use a link line of any length and to set up the coupling system so that power can be transferred adequately over a relatively wide band of frequencies without readjustment of the tuning.

As explained in the chapter on transmitters, the grid circuit of a power amplifier is equivalent to a resistance. The value of the resistance depends on the operating conditions and can be calculated as described in the transmitting chapter. The essentials of the circuit (omitting bias and similar details) for both single-ended and push-pull grids are shown in Fig. 13-22. It is advisable to use coaxial cable for the link, because the shielded construction not only prevents radiation but also makes it possible to install the line in any way that is convenient, without danger of unwanted coupling to other circuits.

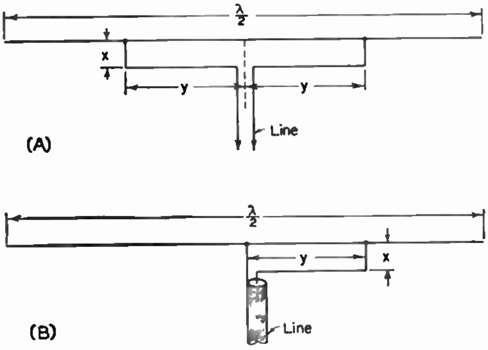


Fig. 13-20 — The "T" match and "gamma" match. The method of adjustment is discussed in the text.

In this system the object of adjustment is to make the s.w.r. in the line as low as possible over as wide a band of frequencies as possible. The controlling factors are the *Q* of the tuned grid circuit, L_1C_1 (see transmitting chapter), the inductance of the coupling coil, L_2, and the degree of coupling between L_1 and L_2. Variable coupling between the coils is convenient, but not strictly necessary if one or both of the other factors can be varied. An s.w.r. indicator (shown as "SWR" in the drawing) is essential. An indicator such as the "Micromatch" (a commercially available instrument) may be connected as shown and the adjustments made under actual operating conditions; that is, with full power applied to the amplifier grid or grids.

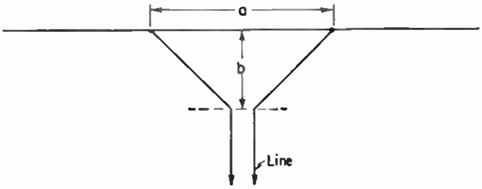


Fig. 13-21 — The "delta" matching section.

Assuming that the coupling is adjustable, start with a trial position of L_2 with respect to L_1, and adjust C_1 for the lowest s.w.r. Then change the coupling slightly and repeat. Continue until the s.w.r. is as low as possible; if the circuit constants are in the right region it should not be difficult to get the s.w.r. down to 1 to 1. The *Q* of the tuned grid circuit should be designed to be at least 10, and if it is not possible to get a very low s.w.r. with such a grid circuit the probable reason is that L_2 is too small. Maximum coupling, for a given degree of physical coupling between the two coils, will occur when the inductance of L_2 is such that its reactance at the operating frequency is equal to the characteristic impedance of the link line. The reactance can be calculated as described in the chapter on electrical fundamentals if the inductance is known; the inductance can either be calculated from the formula in the same chapter or measured as described in the chapter on measurements.

Once the s.w.r. has been brought down to 1 to 1, the frequency should be shifted over the band so that the variation in s.w.r. can be observed, without changing C_1 or the coupling between L_1 and L_2. If the s.w.r. rises rapidly on either side of the original frequency the circuit can be made "flatter" by reducing the *Q* of the tuned grid circuit. This may be done by decreasing C_1 and correspondingly increasing L_1 to maintain resonance, and by tightening the coupling between L_1 and L_2, going through the same adjustment process again. It is possible to set up the system so that the s.w.r. will not exceed 1.5 to 1 over, for example, the entire 7-Mc. band and proportionately on other bands. Under these circumstances a single setting will serve for work anywhere in the band, with essentially constant power transfer from the line to the power-amplifier grids.

If the coupling between L_1 and L_2 is not adjustable the same result may be secured by varying the L/C ratio of the tuned grid circuit — that is, by varying its Q. If any difficulty is encountered it can be overcome by changing the number of turns in L_2 until a match is secured. The two coils should be tightly coupled.

When a resistance-bridge type s.w.r. indicator is used it is not possible to put the full power through the line when making adjustments. In such case the operating conditions in the amplifier grid circuit can be simulated by using a *carbon* resistor, (½ or 1 watt size) of the same value as the calculated amplifier grid impedance, connected as indicated by the arrows in Fig. 13-22. In this case the amplifier tubes *must* be operated "cold" — without filament or heater power. The adjustment process is the same as described above, but with the driver power reduced to a value suitable for operating the s.w.r. bridge.

When the grid coupling system has been adjusted so that the s.w.r. is close to 1 to 1 over the desired frequency range, it is certain that the power put into the link line will be delivered to the grid circuit. The remaining problem, that of coupling power into the line, is discussed in a subsequent section.

Coupling to a Receiver

A good match between an antenna and its transmission line does not guarantee a low standing-wave ratio on the line when the antenna system is used for receiving. The s.w.r. is determined wholly by what the line "sees" at the receiver's antenna-input terminals. For minimum s.w.r. the receiver input circuit must be matched to the line. The rated input impedance of a receiver is a nominal value that varies over a considerable range with frequency. Methods for bringing about a proper match are discussed in the chapter on receivers.

It should be noted that *if* the receiver is matched to the line, then it is desirable that the antenna and line also be matched, since this results in maximum signal transfer from the antenna to the line. If the receiver is *not* matched to the line, the input impedance of the line (at the terminals of the antenna itself) in turn cannot match the antenna impedance. In such a case the signal input to the receiver depends on the coupling system used between the line and the receiver. For greatest signal strength the coupling system has to be adjusted to the best compromise between receiver input impedance and load appearing at the input (antenna) end of the line. The coupling system and its adjustment are a matter for trial and error.

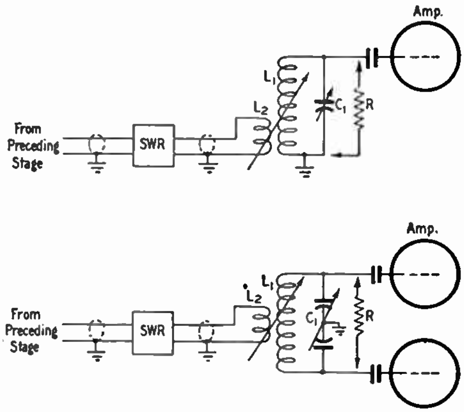


Fig. 13-22 — Set-up for matching a power amplifier grid circuit to a link line. Circuit values are similar to those discussed in the chapter on transmitters. The resistors are used to simulate the grid impedance of the amplifier when a low-power s.w.r. indicator such as a resistance bridge is used. See text for details.

A similar situation exists when the receiver input impedance inherently matches the line Z_0, but the line and antenna are mismatched. Under these conditions perfect matching at the receiver does not result in greatest signal strength; a deliberate mismatch has to be introduced so that the maximum power will be taken from the antenna. Maximum signal input to the receiver occurs when the line is matched at both the receiver and antenna.

Coupling the Transmitter to the Line

In the chapter on transmitters it was explained that an r.f. power amplifier (or oscillator) requires a definite value of load resistance if the desired power output is to be obtained. The present chapter has described the variations in the input impedance of a transmission line with changes in the termination (or s.w.r.) and the line length. Since the input impedance of the line will seldom, if ever, be the same as the load impedance required by the transmitting tube or tubes, it must be transformed to the proper value. Performing this impedance transformation is a primary function of the coupling circuit.

The proper value of load resistance for the amplifier is automatically obtained when the coupling to the line is such that the amplifier, with its plate circuit tuned to resonance, draws the desired plate current. The coupling circuit can be considered to be adjusted satisfactorily when this condition is reached.

Coupling Circuit Requirements

Power tubes must work into a purely resistive load in order to operate at optimum efficiency. On the other hand, the input impedance of a line frequently is partly resistive and partly reactive. In performing the required impedance transformation the coupling circuit must eliminate the reactive part, if any, of the line's input impedance.

In addition, it is highly desirable that the coupling circuit help in preventing any energy on other frequencies than the desired one from reaching the antenna. Although the power in such "spurious" frequencies is comparatively small, interference can be caused to the operation of other services. The coupling circuit should, in other words, have some selectivity.

In the circuits to be discussed, a single-ended tank circuit has been indicated for the amplifier from which power is to be coupled. This is merely to simplify the drawings; any type of amplifier circuit may be substituted. Also, the discussions have been based on the normal operation of the circuits, neglecting such things as stray capacitive coupling which, while having little or no bearing on the fundamental problem of transferring power from one circuit to another, may be very important in causing interference to other services. They are given detailed attention in the chapter on BCI and TVI.

● COUPLING TO FLAT LINES

If the standing-wave ratio on the line is known to be below about 1.5 to 1, the input impedance of the line can be considered to be essentially a pure resistance and a simple transformer can be used to couple the amplifier to the line. As shown in Fig. 13-23, the line is merely connected to a pick-up coil or link coil inductively coupled to the amplifier tank circuit.

Four requirements must be met if the coupling is to be sufficient to load the amplifier properly regardless of line length:

1) The s.w.r. on the line actually must be low.

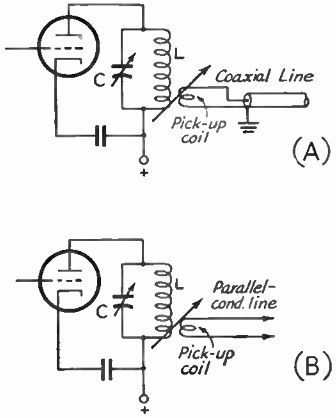


Fig. 13-23 — Using an untuned pick-up coil to couple to a transmission line. The method of adjustment is discussed in the text.

2) The plate tank circuit must have reasonably high Q. A value of 10 or more is usually sufficient.

3) The inductance of the pick-up or link coil must be close to the optimum value for the frequency and type of line used. The optimum coil is one whose self-inductance is such that its reactance at the operating frequency is equal to the Z_0 of the line.

4) It must be possible to make the coupling between the tank and pick-up coils very tight.

The third in this list is often hard to meet, especially when a parallel-conductor line is used. Few, if any, manufactured link coils have adequate inductance for coupling to lines having characteristic impedances over 100 ohms or so, at low frequencies, and very few are large enough even for coupling to 50-ohm line. The optimum pick-up coil for coupling to lines of 300 ohms Z_0 or more will have about the same inductance as the plate tank coil itself.

Amateurs are frequently successful in coupling power into a line even though the pick-up coil is quite small and is loosely coupled to the amplifier tank coil. When such coupling is possible it is an indication that the line is operating at a fairly high s.w.r. and that the line length is such as to bring a current loop near the input end. It is customary to "prune" the line length in such cases until adequate coupling is secured — a practice that has given rise to wholly fallacious beliefs, on the part of many, that pruning the line reduces the standing-wave ratio and that a flat line will load an amplifier with a small link and very loose coupling. Pruning the line accomplishes nothing if the line is actually flat because, as explained earlier in this chapter, the input impedance of a matched line is equal to its Z_0 regardless of the line length. If the line is not flat, pruning changes the input impedance and eventually results in a value such that the link or pick-up coil is actually tuned to the operating frequency by the line, a condition that will give maximum power transfer with minimum coupling. The higher the s.w.r. the more loose the coupling can be.

If the line is operating with a low s.w.r., the systems shown in Fig. 13-23 will require tight coupling between the two coils. Since the secondary (pick-up coil) circuit is not resonant, the leakage reactance of the pick-up coil also will cause some detuning of the amplifier tank circuit. This detuning effect increases with increasing coupling, but is usually not serious. However, the amplifier tuning must be adjusted to resonance, as indicated by the plate-current dip, each time the coupling is changed.

The coupling methods shown in Fig. 13-23, while widely used because of their simplicity, add nothing to the selectivity of the system as a whole and therefore do not contribute to the suppression of spurious radiations.

Tuned Coupling

The design difficulties of using "untuned" pick-up coils, mentioned above, can be avoided by using a coupling circuit tuned to the operating

Fig. 13-24 — Methods for coupling to transmission lines operating at a low standing-wave ratio. The tuned input circuit is frequently essential for obtaining sufficient power transfer, and offers a convenient means for varying loading.

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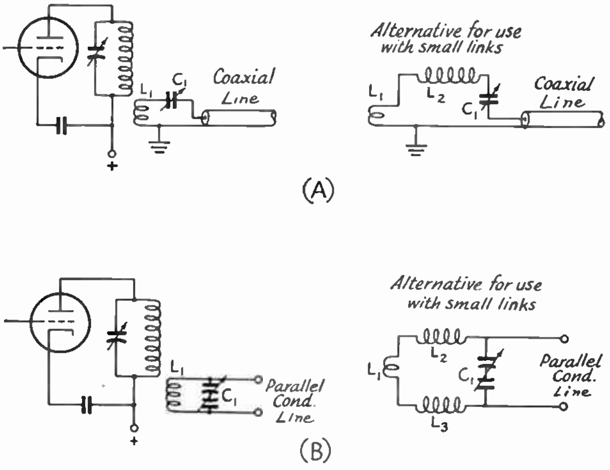

frequency. This contributes additional selectivity as well, and hence aids in the suppression of spurious radiations.

If the line is flat the input impedance will be essentially resistive and equal to the Z_0 of the line. This resistance will be introduced into the coupling circuit, and since the object of tuning is to obtain a reasonable value of Q and thus permit relatively loose coupling, the type of line used will determine the kind of tuning that is practicable. If the line Z_0 is less than 100 ohms, a circuit of reasonable Q can be obtained with practicable values of inductance and capacitance connected in series with the line's input terminals, but connecting the line in parallel with the tuned circuit would require impracticable values of L and C (see chapter on circuit fundamentals). On the other hand, lines having a Z_0 of 300 ohms or more are best connected in parallel with the tuned circuit, since the series connection would require impracticable values of L and C in many cases.

Suitable circuits are given in Fig. 13-24. The values of inductance and capacitance in the coupling circuits are not highly critical, but the L/C ratio must not be too small in the case of series tuning nor too large in the case of parallel tuning. The Q of the coupling circuit often may be as low as 2, without running into difficulty in getting adequate coupling to a tank circuit of proper design. Larger values of Q can be used and will result in increased ease of coupling, but as the Q is increased the frequency range over which the circuit will operate without readjustment becomes smaller. It is usually good practice, therefore, to use a coupling-circuit Q just low enough to permit operation, over as much of a band as is normally used for a particular type of communication, without requiring retuning.

Capacitance values for a Q of 2 and various popular line impedances are given in Table 13-II. In the case of series tuning these are the *maximum* values that should be used; in the case of parallel tuning they are *minimum* values. In both cases the inductance in the circuit should be adjusted to give resonance at the operating frequency. If the link coil used for a particular band does not have enough inductance to resonate, the additional inductance required may be connected in series with it as shown by the alternative circuits.

In practice, the amount of inductance in the circuit should be regulated so that, with somewhat loose coupling between L_1 and the amplifier tank coil, the amplifier plate current will increase when the variable condenser, C_1, is tuned through the value of capacitance given by the table. The coupling between the two coils should then be increased until the amplifier loads normally at the same setting of C_1. Slight retuning of the plate tank condenser may be required. If the transmission line is flat over the entire frequency band under consideration, it should not be necessary to readjust C_1 when changing frequency, if the values given in the table are used. However, it is unlikely that the transmission line actually will be flat over such a range, except possibly with a simple folded dipole antenna, so some readjustment of C_1 may be needed to compensate for changes in the input impedance of the line as the frequency is changed. If the input impedance variations are not large, C_1 may be used as a loading control, no changes in the coupling between L_1 and the tank coil being necessary.

The degree of coupling between L_1 and the amplifier tank coil will depend on the coupling-circuit Q. With a Q of 2, the coupling should be tight — comparable with the coupling that is typical of "fixed-link" manufactured coils. With a swinging link it may be necessary to increase the Q of the coupling circuit in order to get sufficient power transfer. With series tuning this can be done by increasing the L/C ratio; with parallel tuning by decreasing the L/C ratio.

TABLE 13-II

Capacitance in μμfd. Required for Coupling to Flat Lines with Tuned Coupling Circuit

Frequency Band	*Characteristic Impedance of Line*			
Mc.	*52 ohms* [1]	*75 ohms* [1]	*300 ohms* [2]	*600 ohms* [2]
1.8	900	600	600	300
3.5	450	300	300	150
7	230	150	150	75
14	115	75	75	40
28	60	40	40	20

[1] Series tuning: capacitance values are maximum usable.

[2] Parallel tuning: capacitance values are minimum usable.

Note: Inductance in circuit must be adjusted to resonate at operating frequency.

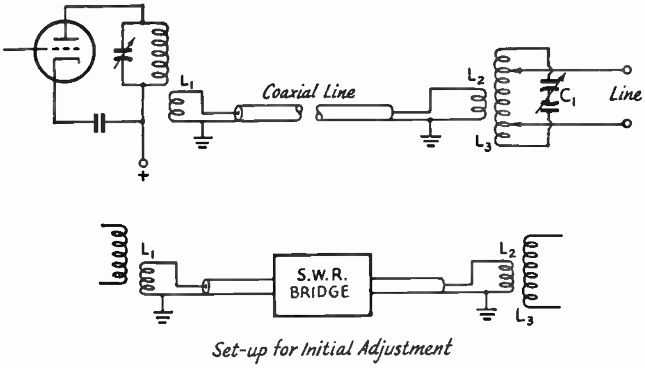

Fig. 13-25 — Matching circuit using a coaxial link, for use with parallel-conductor transmission lines. Adjustment set-up using an s.w.r. bridge is shown in the lower drawing. Design considerations and method of adjustment are discussed in the text.

◆

Link Coupling

The coupling arrangements for parallel-conductor line shown in Fig. 13-24B are not entirely satisfactory from a constructional standpoint. It is usually more convenient to build the coupling apparatus separate from the final amplifier, and this leads to greater operating flexibility as well. For lines operating at a low standing-wave ratio this is easily accomplished by connecting the amplifier and coupling circuits through a short length of transmission line or "link". When properly designed and adjusted, the tuning of both circuits will be completely independent of the length of the line connecting them. This method has the further advantage that, if the connecting line is coaxial cable, it offers an ideal spot for the insertion of low-pass filters for preventing harmonic interference to television and FM reception.

The circuit for coax-link coupling is given in Fig. 13-25. The constants of the tuned circuit C_1L_3 are not particularly critical; the principal requirement is that it must be capable of being tuned to the operating frequency, and constants similar to those used in the plate tank circuit will be satisfactory. The construction of L_3 must be such that it can be tapped at least every turn. L_2 must be tightly coupled to L_3, and the inductance of L_2 should be approximately the value that gives a reactance equal to the Z_0 of the connecting line at the frequency in use. An average reactance of about 60 ohms will suffice for either 52- or 75-ohm coaxial line.

At the amplifier end of the link, L_1 should be as described earlier (Fig. 13-23A or 13-24A) for coupling into coaxial line. The adjustment of coupling at the output end ($L_2L_3C_1$) is entirely independent of the adjustment at the input end (tank circuit and L_1) when the system is properly designed and operated. That is, the circuit formed by $L_2L_3C_1$ acts purely as a matching device to transform the input impedance of the main transmission line to a value equal to the Z_0 of the coaxial link, and the coupling at the amplifier end is consequently designed and adjusted for working into a flat coaxial line.

The most satisfactory way to set up the system initially is to connect a coaxial s.w.r. bridge in the link as shown in Fig. 13-25. A simple resistance bridge such as is described in the chapter on measurements is perfectly adequate, requiring only that the transmitter output be reduced to a very low value so that the bridge will not be overloaded. Take a trial position of the line taps on L_3, keeping them equidistant from the center of the coil, and adjust C_1 for minimum s.w.r. as indicated by the bridge. If the s.w.r. is not close to 1 to 1, try new tap positions and adjust C_1 again, continuing this procedure until the s.w.r. is practically 1 to 1. The setting of C_1 and the tap positions may then be logged for future reference, since they will not change so long as the antenna system and frequency are not changed. At this point, check the link s.w.r. over the frequency range normally used in that band, without changing the setting of C_1. No readjustment will be required if the s.w.r. does not exceed 1.5 to 1 over the range, but if it goes higher it is advisable to note as many settings of C_1 as may be necessary to keep the s.w.r. below 1.5 to 1 at any part of the band. Changes in the link s.w.r. are caused chiefly by changes in the s.w.r. on the main transmission line with frequency, and relatively little by the coupling circuit itself. A single setting of C_1 at mid-frequency will suffice if the antenna itself is broad-tuning.

If it is impossible to get a 1 to 1 s.w.r. at any settings of the taps or C_1, the main transmission line is not flat. Ordinarily there should be no difficulty if the transmission-line s.w.r. is not more than about 3 to 1, but if the line s.w.r. is higher it may not be possible to bring the link s.w.r. down except by using the methods for reactance compensation described in a subsequent section.

Once the matching circuit is properly adjusted, the s.w.r. bridge may be removed and full power applied to the transmitter. The input should be controlled by the coupling between L_1 and the amplifier tank coil, never by making any changes in the settings of the matching circuit. If the amplifier will not load properly, the methods described above (Fig. 13-23A or 13-24A) should be followed.

It is possible to use a circuit of this type without initially setting it up with the s.w.r. bridge. In such a case it is a matter of cut-and-try until adequate power transfer between the amplifier and main transmission line is secured. However, this method frequently results in a high s.w.r. in the link, with consequent power loss, "hot spots" in the coaxial cable, and tuning that is critical with frequency. The bridge method is simple and gives the optimum operating conditions quickly and with certainty.

● "TUNED" LINES

If the s.w.r. on a transmission line is high enough to cause the input impedance to change appreciably as the applied frequency is varied, the coupling between the transmitter and the line must be changed accordingly if the amplifier is to be loaded adequately. So far as the coupling apparatus is concerned, the principal difference between flat and tuned lines is that the system can be designed for relatively constant impedance for flat lines, but must be capable of coupling into a wide range of impedances if the line is "tuned."

The simple circuits shown in Fig. 13-23 can be used for coupling to a line having a high standing-wave ratio providing the line length is adjusted so there is a current loop near the point where it connects to the pick-up coil. The coupling will be maximum, for a given degree of separation between L_1 and the amplifier tank coil, if the line is pruned to a length such that the input impedance is just sufficiently capacitive to cancel the inductive reactance of the pick-up coil. This can be done by cut-and-try. The higher the s.w.r. on the line the easier it becomes to load the amplifier with loose coupling between the two coils. Whether or not good loading can be obtained over a band of frequencies depends on the characteristics of the antenna system. The sharper the antenna and the higher the line s.w.r. the more difficult it becomes to operate over a band without progressively changing the line length.

Series and Parallel Tuning

Rather than adjusting the line length to fit a given coupling coil, it is more practical to adjust the coupling circuit to fit the conditions existing at the input end of the transmission line. The circuit of Fig. 13-24A will serve for coaxial-type lines in all cases that should be tolerated in practice; that is, the range of input impedances that can be handled by such a series-tuned circuit is as great as should be encountered. If series tuning does not work satisfactorily, the s.w.r. is too high for efficient line operation and steps should be taken to get a better match between the line and antenna.

A high standing-wave ratio occurs principally on parallel-conductor lines, either because no attempt has been made at matching the antenna and the line or because the system is used for multiband operation, which precludes such matching. In the latter case, cutting the line length to a multiple of a quarter wavelength will bring either a current or voltage loop near the input terminals of the transmission line (assuming that the antenna itself is resonant) depending on the termination and the line length. If there is a current loop near the input end the impedance will be lower than the line Z_0; if a voltage loop, the input impedance will be higher than the line Z_0. In both cases the input impedances will be essentially resistive.

Under these conditions the circuit arrangements shown in Fig. 13-26 will work satisfactorily. Series tuning is used when a current loop occurs at the input end of the line; parallel tuning when there is a voltage loop at the input end. In the series case, the circuit formed by L_1, C_1 and C_2 with the line terminals short-circuited should tune to the operating frequency. C_1 and C_2 should be maintained at equal capacitance. In the parallel case, the circuit formed by L_1 and C_1 should tune to resonance with the line terminals open.

The L/C ratio in either circuit depends on the transmission line Z_0 and the standing-wave ratio. With series tuning, a high L/C ratio must be used if the s.w.r. is relatively low and the line Z_0 is high. With parallel tuning, a low L/C ratio must be used if the s.w.r. is relatively low and the transmission-line Z_0 also is low. With either series or parallel tuning the L/C ratio becomes less critical when the s.w.r. is high. As a first approximation, L and C values of the same order as those used in the plate tank circuit may be tried.

To adjust the series-tuned circuit, first couple L_1 loosely to the amplifier tank coil and then vary C_1 and C_2, keeping their capacitances equal, until the setting is found that makes the amplifier plate current kick upward. Keep adjusting the amplifier tank condenser, C, for minimum plate current while this is being done. When the proper settings are found, increase the coupling between the two coils until the amplifier draws normal plate current with C adjusted for minimum. It is unnecessary to readjust C_1 and C_2 when the coupling is increased. Keep the coupling between the coils at the smallest value that will load the amplifier properly. If full loading cannot be obtained with the tightest possible coupling, use a coil of more inductance at L_1.

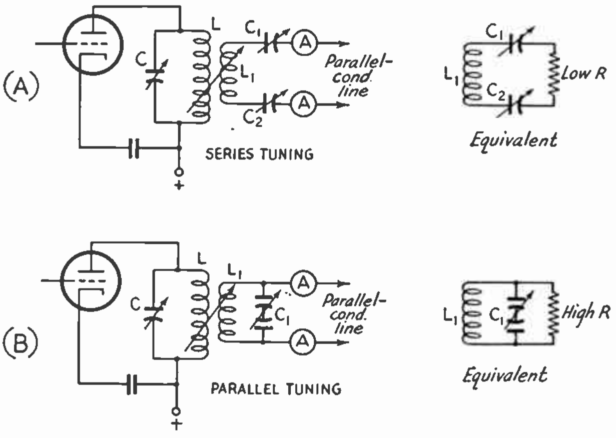

Fig. 13-26 — Series and parallel tuning. This method is useful with resonant lines when the length is such as to bring either a current or voltage loop near the input end. Design data and methods of adjustment are given in the text.

The same adjustment procedure is used with parallel tuning, except that there is only one condenser, C_1. If full loading cannot be secured, reduce the inductance of L_1 and increase C_1 correspondingly to maintain the same frequency.

The r.f. ammeters shown in Fig. 13-26 are not strictly necessary, but are useful for indicating maximum output. They may be omitted if desired; in most cases the amplifier plate current is a good enough indication of output, providing the amplifier is operating at normal ratings and efficiency.

In case full loading cannot be obtained even when the L/C ratio is varied, the type of tuning in use probably is not suitable and should be changed; e.g., from series to parallel. If satisfactory loading still cannot be secured, the probability is that the s.w.r. is quite low and the coupling methods designed for flat lines should be used.

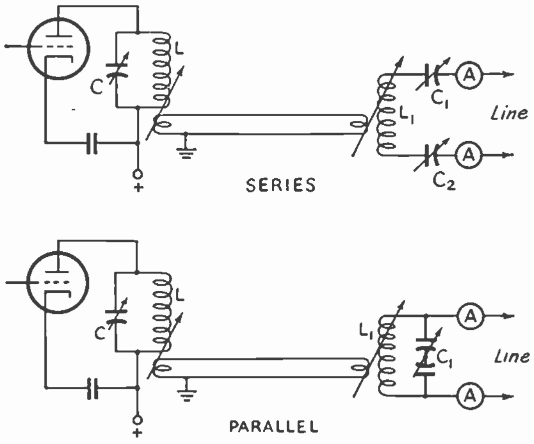

Fig. 13-27 — Link-coupled series and parallel tuning.

Two condensers are used in the series-tuned circuit in order to keep the line balanced to ground. This is because two identical condensers, both connected with either their stators or rotors to the line, will have the same capacitance to ground. A single condenser would be perfectly usable so far as the operation of the coupling circuit is concerned, but will slightly unbalance the circuit, since the frame has more capacitance to ground than the stator. The unbalance is not serious unless the condenser is mounted near a large mass of metal, such as a chassis.

A balanced condenser is used in the parallel circuit, in preference to a single unit, for the same reason. An alternative scheme to maintain balance is to use two single-ended condensers in parallel, but with the frame of one connected to one side of the line and the frame of the other connected to the other side of the line. The same two condensers may be switched in series when series tuning is to be used..

Link Coupling

The circuits shown in Fig. 13-26 require a means for varying the coupling between two sizable coils, a thing that is somewhat inconvenient constructionally. It is easier to use separate fixed mountings for the final tank and antenna coils and couple them by means of a link. As explained in the chapter on circuit fundamentals, a *short* link is equivalent to providing mutual inductance between two tuned circuits. Typical arrangements for series and parallel tuning are shown in Fig. 13-27. Although these drawings show variable coupling at both ends of the link, a fixed link coil can be used at either end so long as variable coupling is available at the other.

There is no essential difference between the tuning procedures with these circuits and those of Fig. 13-26. The only change is that the coupling is adjusted by means of a link instead of by varying the spacing between L and L_1.

In cases where the link will be more than a few inches long, or when coaxial cable is to be used for the link, it is much better to consider the link as a transmission line that should be properly matched. The circuit of Fig. 13–25 is recommended in that case, except that either a series- or parallel-tuned circuit is substituted for C_1L_3 in that figure. The same considerations apply with respect to the sizes of the link coils, and the best adjustment procedure is that using an s.w.r. bridge.

Lines of Random Length

Series or parallel tuning will always work satisfactorily with lines having a high standing-wave ratio so long as the electrical length of the line is approximately a multiple of a quarter wavelength. However, it is not always possible to couple satisfactorily when intermediate line lengths are used. This is because at some lengths the input impedance of the line has a considerable reactive component, and because the resistive component is too large to be connected in series with a tuned circuit and too low to be connected in parallel.

The coupling system shown in Fig. 13-25 is capable of handling the resistive component of the input impedance of the transmission lines used in most amateur installations, regardless of the standing-wave ratio on the line. (A possible exception is where the s.w.r. is considerably higher than 10 to 1 and the line length is such as to bring a current loop at the input end. In such a case the resistance may be only a few ohms, which is difficult to match by means of taps on a coil.) Consequently, it can generally be used wherever either series or parallel tuning would normally be called for, simply by setting the taps properly on the coil.

Within limits, the same circuit is capable of being adjusted to compensate for the reactive component of the input impedance; this merely means that a 1 to 1 s.w.r. in the link will be obtained at a different setting of C_1 (Fig. 13-25) than would be the case if the line "looked like" a pure resistance. Sometimes, however, C_1 does not have enough range available to give complete compensation, particularly when (as is the case with some line lengths when the s.w.r. is high) the input impedance is principally reactive.

Under such conditions it is necessary, if the line length cannot be changed to a more satisfactory value, to provide additional means for compensating for or "cancelling out" the reactive component of the input impedance. As described earlier in this chapter (Fig. 13-6) the input impedance can be considered to be equivalent to a circuit consisting either of resistance and inductance or resistance and capacitance. It is generally more convenient to consider these elements as a parallel combination, so if the line "looks like" $L'R'$ at A in Fig. 13-6, it is apparent that if we connect a capacitance of the right value across L' the circuit will become resonant and will appear to be a pure resistance of the value R'. Similarly, connecting an inductance of the right value across C' in Fig. 13-6B will resonate the circuit and the impedance will be equal to R'. The resistive impedance that remains can easily be matched to the coax link by means of the circuit of Fig. 13-25.

The practical application of this principle is shown in Fig. 13-28, where L and C are the reactances required to cancel out the line reactance, L for cases where the line is capacitive, C for lines having inductive reactance. The amount of either inductance or capacitance required is easily determined by trial. Using the s.w.r. bridge in the coax link, first disconnect the main transmission line and connect a noninductive resistor to the line terminals. A ½ or 1-watt carbon resistor of about the same resistance as the line Z_0 will do. Adjust the coil taps and C_1 for a 1 to 1 standing-wave ratio in the link, as described earlier. This determines the proper setting of C_1 for a purely resistive load. Then take off the resistor and connect the line, again adjusting the taps and C_1 for minimum s.w.r. If a 1 to 1 ratio can be obtained further compensation is not needed, but if not, make the s.w.r. as low as possible and compare the new setting of C_1 with the original setting. If the capacitance has increased, the line reactance is inductive and a condenser must be connected at C in Fig. 13-28. The amount of capacitance needed to bring the proper setting of C_1 near the original setting can be determined by trial. On the other hand, if the capacitance of C_1 is less than the original, an inductance must be connected at L. Trial values will show when the proper tuning conditions have been reached. It is not necessary that C_1 be at exactly the original setting after the compensating reactance has been adjusted; it is sufficient that it be somewhere in the same vicinity.

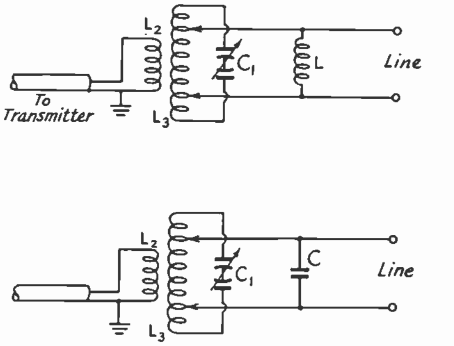

Fig. 13-28 — **Reactance cancellation on random-length lines having a high standing-wave ratio.**

Using this procedure practically any length of line can be coupled properly to the transmitter, even when the line s.w.r. is quite high. Unfortunately, no specific values can be suggested for L and C, since they vary widely with line length and s.w.r. Their values usually are comparable with the values used in the regular coupling circuits at the same frequency.

Coupler or Matching Circuit Construction

The design of matching or "antenna coupler" circuits has been covered in the preceding section, and the adjustment procedure also has been outlined. Since circuits of this type are most frequently used for transferring power from the transmitter to a parallel-conductor transmission line, a principal point requiring attention is that of maintaining good balance to ground. If the coupler circuit is appreciably unbalanced the currents in the two wires of the transmission line will also be unbalanced, resulting in radiation from the line.

In most cases the matching circuit will be built on a metal chassis, following common practice in the construction of transmitting units. The chassis, because of its relatively large area, will tend to establish a "ground" — even though not actually grounded — particularly if it is assembled with other units of the transmitter in a rack or cabinet. The components used in the coupler, therefore, should be placed so that they are electrically symmetrical with respect to the chassis and to each other.

In general, the construction of a coupler circuit should physically resemble the tank layouts used with push-pull amplifiers. In parallel-tuned circuits used with matched lines and with tuned lines having high input impedance, a spilt-stator condenser should be used. The condenser frame should be insulated from the chassis because, depending on line length and other factors, harmonic reduction and line balance may be improved in some cases by grounding and in others by not grounding. It is therefore advisable to adopt construction that permits either. Provision also should be made for grounding the center of the coil, for the same reason. A single coil used in a parallel-tuned circuit should be mounted so that its hot ends are symmetrically placed with respect to the chassis and other components. This equalizes stray capacitances and helps maintain good balance.

When the coupler is of the type that can be shifted to series or parallel tuning as required, two separate single-ended condensers will be satisfactory. As described earlier, they should be connected so that both frames go to the same side of the circuit — i.e., either to the coil or to the line — for series tuning, and should be connected frame-to-stator for parallel tuning.

A coupler designed and adjusted so that the connecting link acts as a matched transmission line may be placed in any convenient location. Some amateurs prefer to install the coupler at the point where the main transmission line enters the station. This helps maintain a neat station layout when an air-insulated parallel-conductor transmission line is used. With solid-dielectric lines, which lend themselves well to neat installation indoors, it is probably more desirable to install the coupler where it can be reached easily for adjustment and band-changing. The use of coaxial line between the transmitter and coupler is strongly recommended if the link line is more than a few inches long, for the reasons outlined in the preceding section.

COAX-COUPLED MATCHING CIRCUIT

The matching unit shown in Figs. 13-29 to 13-32, inclusive, is constructed according to the design principles outlined earlier in this chapter. It uses a parallel-tuned circuit with taps for matching a parallel-conductor line through a link coil to a coaxial line to the transmitter. It will handle about 500 watts of r.f. power and will work, without modification, into lines having an s.w.r. below 3 or 4 to 1. If the s.w.r. is high, it may be necessary to compensate for the reactive part of the input impedance of the line, at certain line lengths, by using an additional coil or condenser as discussed earlier. The necessity for such compensation can be avoided, on lines having a high s.w.r., by making the electrical length of the line a multiple of a quarter wavelength.

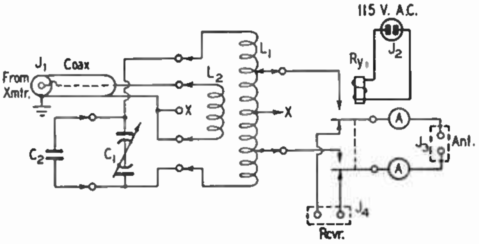

Fig. 13-30 — Circuit diagram of the antenna coupler. The antenna changeover relay and r.f. ammeters are convenient but not essential to the operation of the coupler.

The ground (X in the diagram) on the center of the tank coil may be used or not, as required for best harmonic suppression.

C_1 — 100-μμfd.-per-section, 1500-volt plate spacing per section (National TMK-100D).
C_2 — 90 μμfd., 3000 volts, 2 amp. at 3 Mc., mica.
A — R.f. ammeter, scale range according to power and antenna feeder system. For 300-ohm line operating at less than 3-to-1 s.w.r., 0–1 amp. is satisfactory for 100 watts r.f. output; 0–2 will suffice for outputs up to 400 watts.
J_1 — Coax receptacle.
J_2 — 115-volt receptacle, male (Amphenol).
J_3, J_4 — Crystal socket, for FT-243-type pin spacing (Millen 33102).
Ry_1 — Antenna relay, d.p.d.t. (Ward Leonard 507–531).

Coil Data

Band	*L_1, turns*	*L_2, turns*
3.5–4 Mc.	24	8
7 Mc.	18	5
14 Mc.	10	3
28 Mc.	6	2

L_1 — No. 12 tinned wire, 2½ inches dia., 6 turns per inch (B & W 3905–1).
L_2 — No. 14 tinned wire, 2 inches dia., 8 turns per inch (B & W 3900).

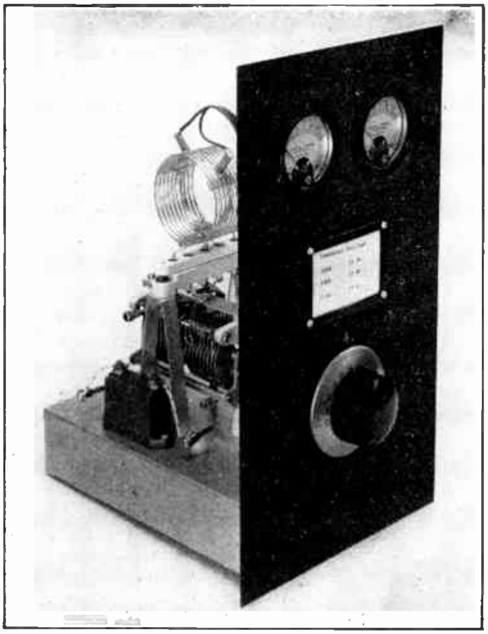

Fig. 13-29 — A matching circuit or "antenna coupler" for use between a coaxial link line and a parallel-conductor transmission line. Link coil design is optimum for sufficient power transfer from a flat coaxial line.

As shown by Fig. 13-30, the circuit includes an antenna changeover relay and r.f. ammeters for measuring feeder current. Neither is essential to the operation of the coupler, but they are frequently used and the photographs show how they may be incorporated in the coupler unit. The coupler is fitted with crystal sockets for plugs such as the Millen type 37412 for 300-ohm line; the plug-and-socket arrangement facilitates shifting antennas, in case two or more are available for different bands. The same plug-and-socket combination may be used with lines of other characteristic impedances, but other types of plugs and sockets, or binding posts, may be substituted.

The coils are constructed from commercial coil material to meet the link inductance requirements outlined earlier in this chapter. The diameter of the link coil is such that it fits snugly inside the tank coil, and once the coils are cut to the proper size they may be cemented together

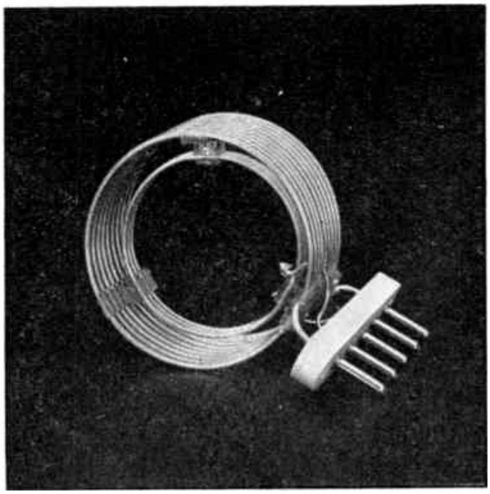

Fig. 13-31 — **Construction of coils used in the antenna coupler. Connections to the link coil are made with short lengths of wire soldered to the ends of the link winding. The tank-coil turns should be spread slightly where the connecting wires to the link pass through the coil, to prevent short circuits. The link connections should be run at right angles to the tank coil turns, insofar as possible, to reduce capacitive coupling.**

at their tie-strips, using Duco cement. A typical coil assembly is shown in Fig. 13-31, the coils being mounted on Millen type 40305 plugs and requiring no other support than the stiffness of the short lengths of wire at the ends where they go into the prongs on the plug. The taps on the tank coil for matching are made by means of Johnson type 235-860 clips.

The coupler is built on a 7 by 9 by 2 aluminum chassis. The coil socket (Millen 41305) is mounted on brackets made from 1/16-inch aluminum cut in strips a half inch wide, just high enough so that the coil socket clears the tuning condenser comfortably. A vernier-type friction-drive dial (National AM) is used on the tuning condenser. The r.f. meters are mounted on a piece of bakelite set behind a rectangular cut-out in the 8 by 12 metal panel, the bakelite being used to reduce the capacitance between the meters and the panel.

Each coil is provided with its own pair of clips soldered to a short length of 300-ohm line terminated in a plug. The line plug is inserted in a crystal socket mounted on top of the tuning condenser. This method avoids the necessity for changing clip connections when changing coils.

The socket for the coax link is mounted centrally on the rear chassis edge. A length of coax runs from this coax receptacle to the coil socket, and is grounded where it goes through the chassis (between the stator sections of the tuning condenser) to reach the coil socket. The a.c. plug on the rear edge connects to the relay coil.

A fixed condenser, C_2, can be connected in parallel with C_1 by flexible leads and banana plugs. The sockets (taken from jack-top binding posts) are soldered to the lugs on the variable condenser. C_2 is used for padding the circuit on the 3.5-Mc. range only; a sufficiently large coil for tuning with the 50 $\mu\mu$fd. available in C_1 cannot be mounted on the plug bar.

The *L/C* ratio in the coupler tank circuit is not especially critical, so the dimensions given in Fig. 13-30 can be varied within reason. The chief point is that each coil must resonate, on the band for which it is made, with the tuning condenser. Fairly low *C* is preferable to high *C*, but the limitation on *L/C* ratio at 3.5 Mc. is the size of the coil required. The plug bar will mount a 4-inch-long coil comfortably. With the 3.5-Mc. coil dimensions given in Fig. 13-30 the coil is just slightly longer than the bar and is easily mounted. Additional support is given this coil by running a No. 4 screw through the end holes in the bar and fastening a soldering lug under the nut; the coil ends are soldered to this lug as well as to the pins in the plug.

The link coils specified have adequate inductance for full coupling with either 52- or 75-ohm coaxial link lines.

The preferable method of adjusting the coupler is that using an s.w.r. bridge, designed for the characteristic impedance of the coaxial line used for the link, as described in a preceding section in this chapter.

● SERIES-PARALLEL COUPLER FOR WALL MOUNTING

Fig. 13-33 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

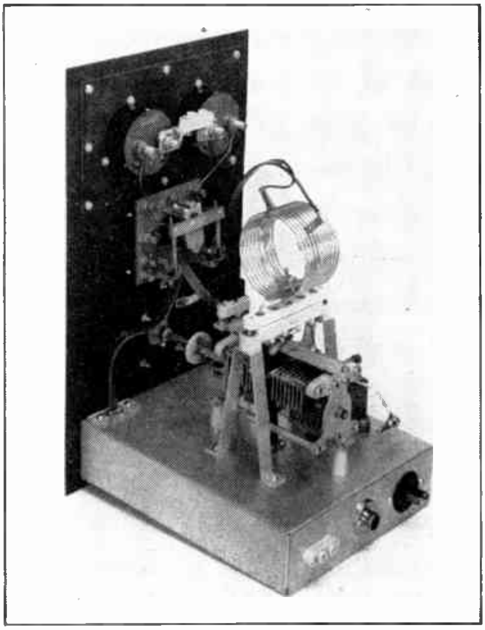

Fig. 13-32 — **Rear view of the antenna coupler. Connections to the coax link, to the receiver antenna posts, and to the 115-volt supply for the relay are through sockets on the rear edge of the chassis.**

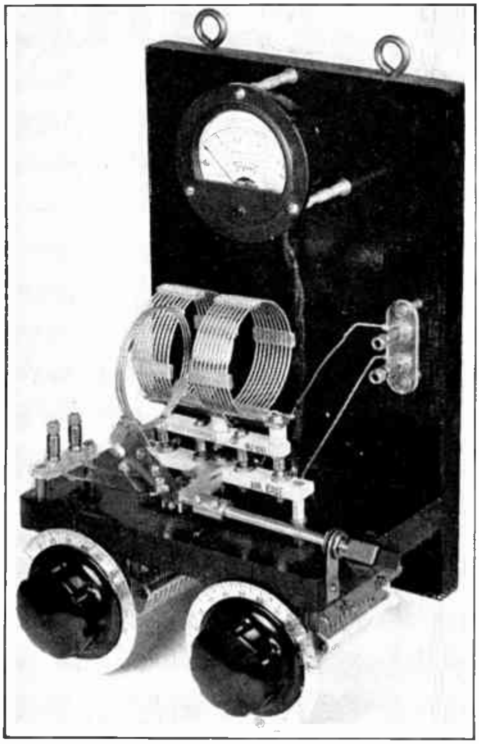

Fig. 13-33 — A wall-mounting antenna coupler for medium-power transmitters. This unit provides a choice of either series or parallel tuning for resonant feeders. Standard transmitting coils of the variable-link type are used.

using a similar layout, but substituting heavier coils and condensers with greater plate spacing.

As shown in Fig. 13-34, 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 × 7/8-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.

The variable-link windings of manufactured coils may not give adequate coupling unless the length of the link-line to the transmitter is adjusted, by cut-and-try, for optimum results. As an alternative, the coil sets may be wound as described for the antenna coupler shown in Fig. 13-29. The coils dimensions given will be satisfactory for use in the circuit of Fig. 13-34, and providing a coaxial link is used the coupling will be independent of link-line length when adjusted by means of an s.w.r. bridge.

● RACK-MOUNTING SERIES-PARALLEL COUPLER

The rack-mounting coupling unit shown in Fig. 13-35 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 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.

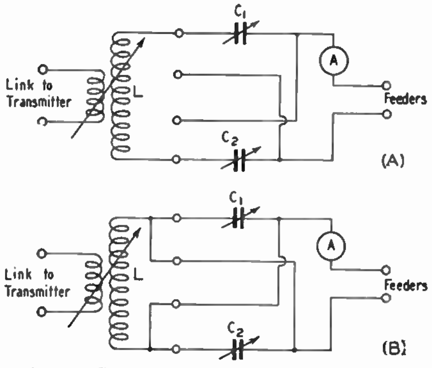

Fig. 13-34 — Circuit diagram of an antenna coupler for use with a medium-power transmitter. A — Series tuning. B — Parallel tuning.

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.

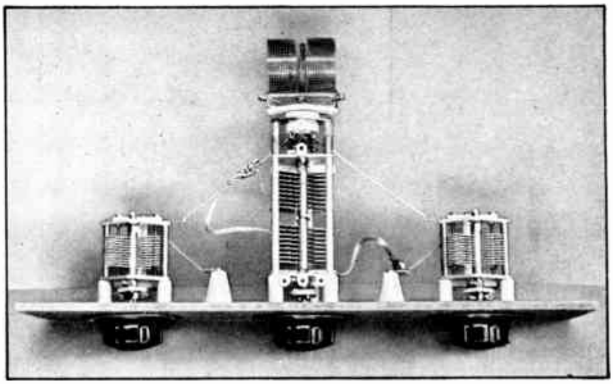

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Fig. 13-35 — Rack-mounted coupler for low-power transmitters. This unit uses three variable condensers to provide either series or parallel tuning without condenser switching.

◆

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. 13-36. 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.

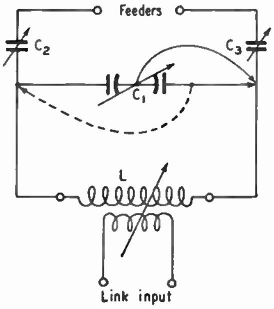

Fig. 13-36 — Circuit of the rack-mounting antenna tuner for use with transmitters having final amplifiers that are operated at less than 1000 volts on the plate.

All coils are $1\frac{7}{8}$ inches in diameter and $2\frac{1}{4}$ 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 $\mu\mu$fd. 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 $\mu\mu$fd., 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.

● A WIDE-RANGE ANTENNA COUPLER

The photograph of Fig. 13-37 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. 13-38.

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 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. 13-38.

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

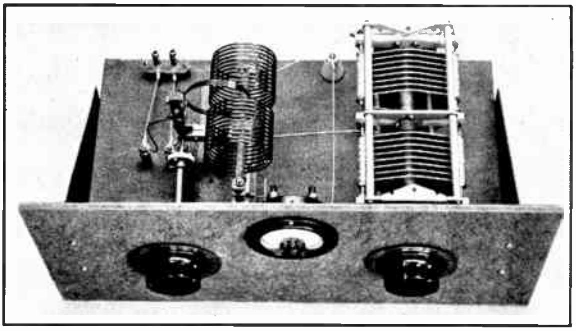

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Fig. 13-37 — Wide-range antenna coupler. The unit is assembled on a metal chassis measuring 10 × 17 × 2 inches, with a panel 8¾ × 19 inches in size. The variable condenser is a split-stator unit with a capacitance of 200 $\mu\mu$fd. per section and 0.07-inch plate spacing (Johnson 200ED30). The plug-in coils are the B & W TVL series. The r.f. ammeter has a 4-ampere scale.

◆

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.

In case there is difficulty in loading the transmitter with particular lengths of link line, adequate coupling usually can be secured by tuning the link circuit as described earlier in this chapter. On the lower frequencies it may be necessary to add inductance in series with the link coil in order to raise the *Q* of the link circuit to a sufficiently high value for good coupling.

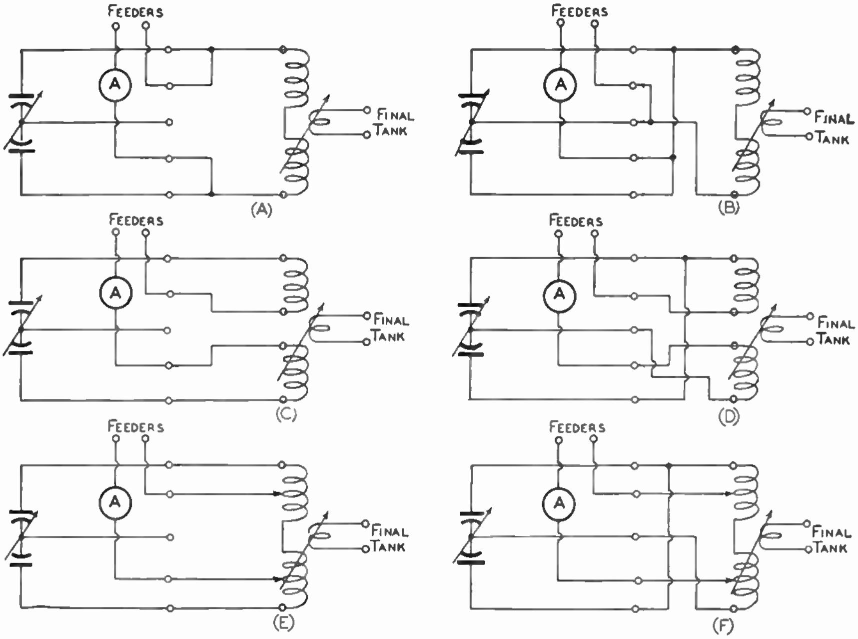

Fig. 13-38 — Circuit diagram of the wide-range rack-type antenna coupler. A — Parallel tuning, low *C*. B — Parallel tuning, high *C*. C — Series tuning, low *C*. D — Series tuning, high *C*. E — Parallel tank, low-impedance output, low *C*. F — Parallel tank, low-impedance output, high *C*. After the 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

An *antenna system* can be considered to include the antenna proper (the portion that radiates the r.f. energy), the feedline, and any coupling devices used for transferring power from the transmitter to the line and from the line to the antenna. Some simple systems may omit the transmission line or one or both of the coupling devices. This chapter will describe the antenna proper, and in many cases will show popular types of lines, as well as line-to-antenna couplings where they are required. However, it should be kept in mind that *any* antenna proper can be used with *any* type of feedline if a suitable coupling is used between the antenna and the line. Changing the line does not change the type of antenna.

Selecting an Antenna

In selecting the type of antenna to use, the majority of amateurs are somewhat limited through space and structural limitations to simple antenna systems, except for v.h.f. operation where the small space requirements make the use of multielement beams readily possible. This chapter will consider antennas for frequencies as high as 30 Mc. — a later chapter will describe the popular types of v.h.f. antennas. However, even though the available space may be limited, it is well to consider the propagation characteristics of the frequency band or bands to be used, to insure that best possible use is made of the available facilities. The propagation characteristics of the various bands, up to 30 Mc., are described in Chapter Four. In general, antenna construction and location become more critical and important on the higher frequencies. On the lower frequencies (3.5 and 7 Mc.) the vertical angle of radiation and the plane of polarization may be of relatively little importance; at 28 Mc. they may be all-important. On a given frequency, the particular type of antenna best suited for long-distance communication may not be as good for shorter-range work as would a different type.

Definitions

The important properties of an antenna proper are its polarization, vertical and horizontal angles of maximum radiation, impedance, gain and bandwidth.

The **polarization** of a straight-wire antenna is determined by its position with respect to the earth. Thus a vertical antenna radiates vertically-polarized waves, while a horizontal antenna radiates horizontally-polarized waves in a direction broadside to the wire and vertically-polarized waves at high vertical angles off the ends of the wire. The wave from an antenna in a slanting position, or from the horizontal antenna in directions other than mentioned above, contains both horizontal and vertical components.

The **vertical angle of maximum radiation** of an antenna is determined by the free-space pattern of the antenna, its height above ground, and the nature of the ground. The angle is measured in a vertical plane with respect to a tangent to the earth at that point, and it will usually vary with the horizontal angle, except in the case of a simple vertical antenna. The **horizontal angle of maximum radiation** of an antenna is determined by the free-space pattern of the antenna.

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 to the line offered by the antenna. It can be either resistive or complex, depending upon whether or not the antenna is resonant.

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.

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.

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.

The **bandwidth** of an antenna generally refers to the frequency range over which the gain and impedance are substantially constant. It is of importance primarily in connection with multielement beams fed by a "flat" transmission line.

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 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.

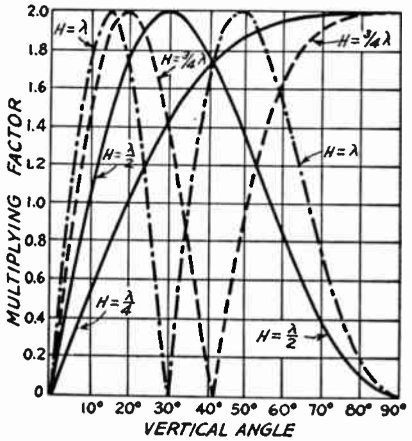

Fig. 14-1 — **Effect of ground on radiation of horizontal antennas at vertical angles for four antenna heights. This chart is based on perfectly-conducting ground.**

Fig. 14-1 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 of maximum 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 angles usually are most effective, this generally means that the antenna should be high — at least one-half wavelength at 14 Mc., and preferably three-quarters or one wavelength, and at least one wavelength, and preferably higher, at 28 Mc. 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 useful 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. 14-1 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 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. 14-2. 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 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.

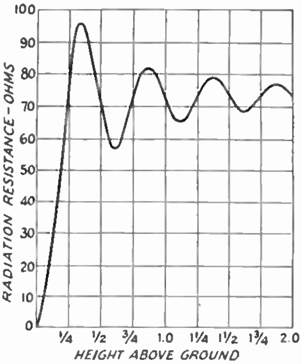

Fig. 14-2 — Theoretical curve of variation of radiation resistance for a half-wave horizontal antenna, as a function of height in wavelength above perfectly-reflecting ground.

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:

$$\textit{Length}\text{ (feet)} = \frac{492}{\textit{Freq.}\text{ (Mc.)}} \qquad \textbf{(14-A)}$$

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. 14-3, 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.:

$$\textit{Length of half-wave antenna}\text{ (feet)} = \frac{492 \times 0.95}{\textit{Freq.}\text{ (Mc.)}} = \frac{468}{\textit{Freq.}\text{ (Mc.)}} \qquad \textbf{(14-B)}$$

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. 14-3.

$$\textit{Length of half-wave antenna}\text{ (feet)} = \frac{492 \times K}{\textit{Freq.}\text{ (Mc.)}} \qquad \textbf{(14-C)}$$

$$\textit{or length}\text{ (inches)} = \frac{5905 \times K}{\textit{Freq.}\text{ (Mc.)}} \qquad \textbf{(14-D)}$$

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. 14-A. Ratio of half-wavelength to conductor diameter (changing wavelength to inches) is $\frac{16.97 \times 12}{2} = 101.8$. From Fig. 14-3, $K = 0.963$ for this ratio. The length of the antenna, from Eq. 14-C, 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. 14-D: $\frac{5905 \times 0.963}{29} = 196$ inches.

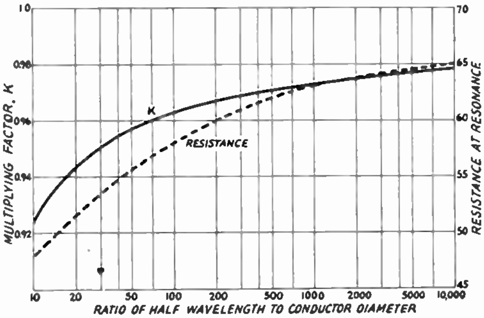

Fig. 14-3 — Effect of antenna diameter on length for half-wave resonance, shown as a multiplying factor, K, to be applied to the free-space half-wavelength (Equation 14-A). The effect of conductor diameter on the impedance measured at the center also is shown.

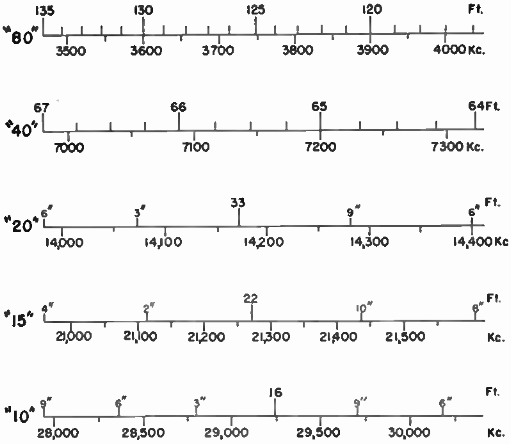

Fig. 14-4 — The above scales, based on Eq. 14-B, can be used to determine the length of a half-wave antenna of wire.

Current and Voltage Distribution

When power is fed to a half-wave 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.

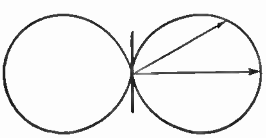

Fig. 14-5 — The free-space radiation pattern of a half-wave antenna. The antenna is shown in the vertical position. This is a cross-section of the solid pattern described by the figure when rotated on its vertical axis. The "doughnut" form of the solid pattern can be more easily visualized by imagining 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.

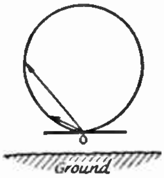

Fig. 14-6 — Illustrating the importance of vertical angle of radiation in determining antenna directional effects. Off the end, the radiation is greater at higher angles. Ground reflection is neglected in this drawing of the free-space pattern of a horizontal antenna.

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 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. 14-3. 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.

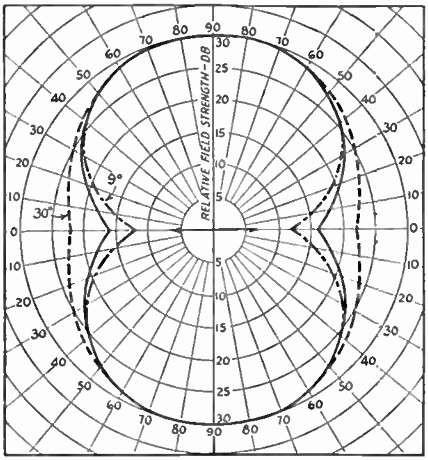

Fig. 14-7 — Horizontal pattern of a horizontal half-wave antenna at three vertical radiation angles. The solid line is relative radiation at 15 degrees. Dotted lines show deviation from the 15-degree pattern for angles of 9 and 30 degrees. The patterns are useful for shape only, since the amplitude will 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.

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. 14-5, 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. 14-6 and 14-7.

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. 14-8. In the method shown at A, C_1 and C_2 should be about 150 μμfd. each for the 3.5-Mc. band, 75 μμfd. each at 7 Mc., and proportionately smaller at the higher frequencies. The antenna coil connected between them should resonate to 3.5 Mc. with about 60 or 70 μμfd., for the 80-meter band, for 40 meters it should resonate with 30 or 35 μμfd., 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.

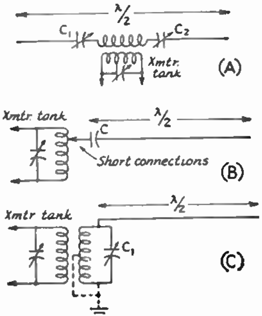

Fig. 14-8 — Methods of directly exciting the half-wave antenna. A, current feed, series tuning; B, voltage feed, capacitive coupling; C, voltage feed, with inductively-coupled antenna tank. In A, the coupling circuit is not included in the effective electrical length of the antenna system proper.

The circuits in Fig. 14-8B and C are used when only one end of the antenna is accessible. In B, the coupling is adjusted by moving the 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.

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 **14-B**, 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. 14-9.

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 dipole,** a good match is offered for a 300-ohm line. Such an antenna is shown in Fig. 14-10. The open-wire line shown in Fig. 14-10 is made of

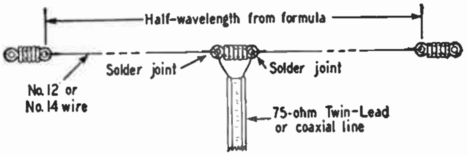

Fig. 14-9 — Construction of a half-wave doublet fed with 75-ohm line. The length of the antenna is calculated from Equation 14-B or Fig. 14-4.

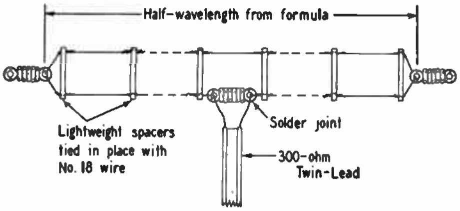

Fig. 14-10 — The construction of an open-wire folded doublet fed with 300-ohm line. The length of the antenna is calculated from Equation 14-B or Fig. 14-4.

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.

The half-wavelength antenna can also be made from the proper length of 300-ohm line, opened on one side in the center and connected to the feedline. After the wires have been soldered together, the joint can be strengthened by molding some of the excess insulating material (polyethylene) around the joint with a hot iron, or a suitable lightweight clamp of two pieces of Lucite can be devised.

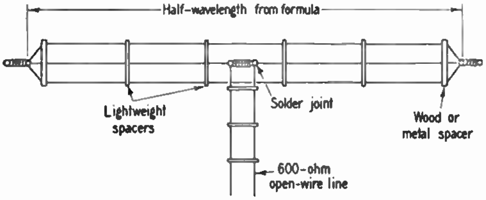

Fig. 14-11 — The construction of a 3-wire folded dipole is similar to that of the 2-wire folded dipole. The end spacers may have to be slightly stronger than the others because of the greater compression force on them. The length of the antenna is obtained from Equation 14-B or Fig. 14-4. A suitable line can be made from No. 14 wire spaced 4½ to 5 inches, or from No. 12 wire spaced 6 inches.

Similar in some respects to the two-wire folded dipole, the three-wire folded dipole of Fig. 14-11 offers a good match for a 600-ohm line. It is favored by amateurs who prefer to use an open-wire transmission line instead of the 300-ohm insulated line. The three wires of the antenna proper should all be of the same diameter.

Another method for offering a match to a 600-ohm open-wire line with a half-wavelength antenna is shown in Fig. 14-12. 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 14-B or Fig. 14-4. The length of section *C* is computed from:

$$C \text{ (feet)} = \frac{118}{Freq. \text{ (Mc.)}} \qquad \textbf{(14-E)}$$

The feeder clearance, *E*, is found from

$$E \text{ (feet)} = \frac{148}{Freq. \text{ (Mc.)}} \qquad \textbf{(14-F)}$$

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.

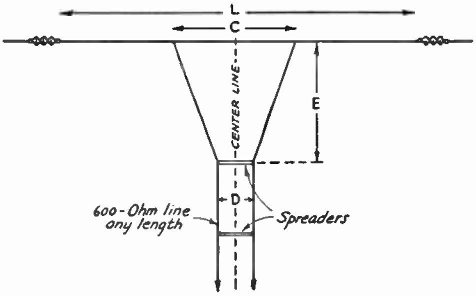

Fig. 14-12 — Delta-matched antenna system. The dimensions *C*, *D*, and *E* are found by formulas given in the text. It is important that the matching section, *E*, come straight away from the antenna without any bends.

Since the equations hold only for 600-ohm line, it is important that the line be close to this value. This requires 4¾-inch spaced No. 14 wire, 6-inch spaced No. 12 wire, or 3¾-inch spaced No. 16 wire.

If a half-wavelength antenna is fed at the center with other than 75-ohm line, or if a two-wire dipole 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 in the preceding 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

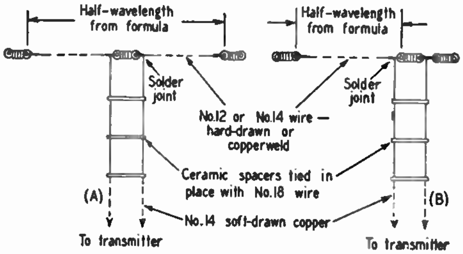

Fig. 14-13 — The half-wave antenna can be fed at the center or at the end with an open-wire line. The antenna length is obtained from Equation 14-B or Fig. 14-4.

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. 14-13. If the power level is low, below 100 watts or so, 300-ohm Twin-Lead can be used in place of the open line.

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. 14-14 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.

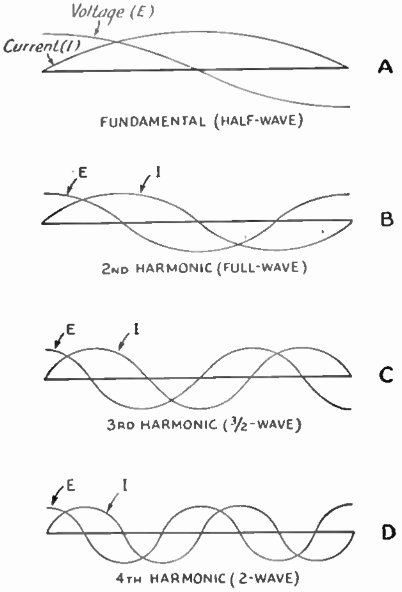

Fig. 14-14 — Standing-wave current and voltage distribution along an antenna when it is operated at various harmonics of its fundamental resonant frequency.

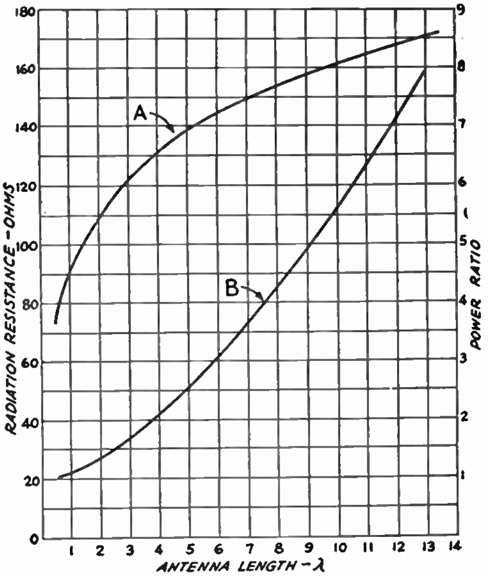

Fig. 14-15 — Curve *A* shows variation in radiation resistance with antenna length. Curve *B* shows power in lobes of maximum radiation for long-wire antennas as a ratio to the maximum radiation for a half-wave antenna.

The polarity of current or voltage in each standing wave is opposite to that in the adjacent 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.

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

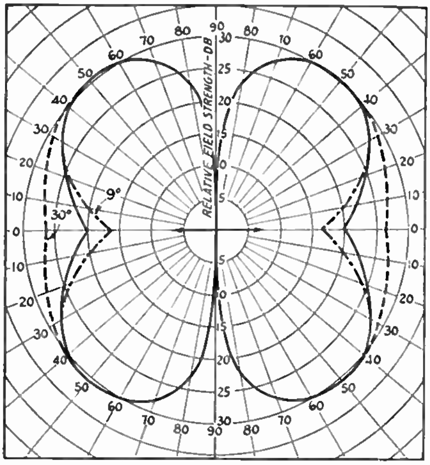

Fig. 14-16 — Horizontal patterns of radiation from a *full-wave* antenna. The solid line shows the pattern for a vertical angle of 15 degrees; dotted lines show deviation from the 15-degree pattern at 9 and 30 degrees. All three patterns are drawn to the same relative scale; actual amplitudes will depend upon the height of the antenna.

portion of the wave in space. The formula for the length of a long-wire antenna, therefore, is

$$Length \text{ (feet)} = \frac{492\ (N-0.05)}{Freq.\ \text{(Mc.)}} \qquad \textbf{14-G}$$

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 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, the frequency for which the antenna is cut.

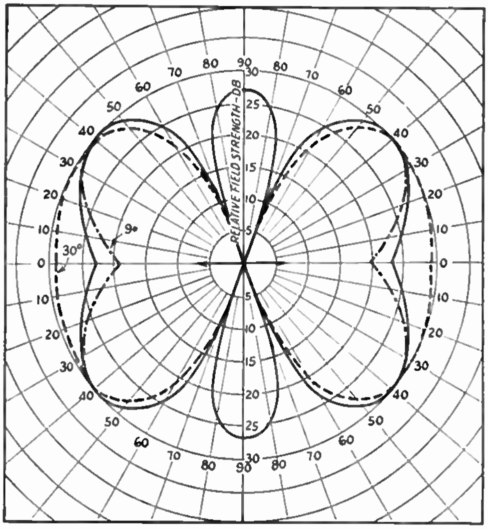

Fig. 14-17 — Horizontal patterns of radiation from an antenna *three half-waves* long. The solid line shows the pattern for a vertical angle of 15 degrees; dotted lines show deviation from the 15-degree pattern at 9 and 30 degrees. Minor lobes coincide for all three angles.

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. 14-15 shows how the radiation resistance and the power in the lobe of maximum radiation vary with the antenna length.

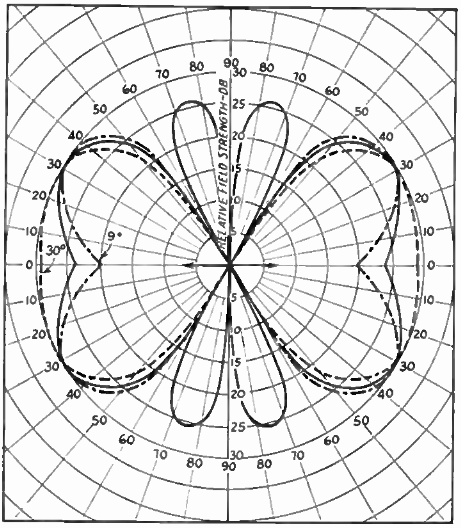

Fig. 14-18 — Horizontal patterns of radiation from an antenna *two wavelengths* long. The solid line shows the pattern for a vertical angle of 15 degrees; dotted lines show deviation from the 15-degree pattern at 9 and 30 degrees. The minor lobes coincide for all three angles.

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. 14-16, 14-17 and 14-18, 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.

Methods of Feeding

In a long-wire antenna, the currents in adjacent half-wave sections must be out of phase, as shown in Fig. 14-14. 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 in the preceding 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 solid-dielectric 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 solid dielectric. It is wise not to attempt to use on its harmonics a half-wave antenna center-fed with coaxial cable. High-impedance solid-dielectric lines such as 300-ohm Twin-Lead may be used, however, provided the power does not exceed a few hundred watts.

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.

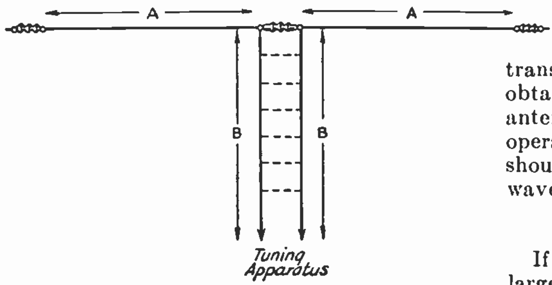

Fig. 14-19 — Practical arrangement of a shortened antenna. The total length, $A + B + B + A$, should be a half-wavelength for the lowest-frequency band, usually 3.5 Mc. See Table 14-I for lengths and tuning data.

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. 14-16, but if it is fed in the center the pattern will be somewhat similar to Fig. 14-7, 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 14-I 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

TABLE 14-I
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.

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 one half-wave around the whole system.

A practical antenna of this type can be made as shown in Fig. 14-19. Table 14-II 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 building 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. 14-20. 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.

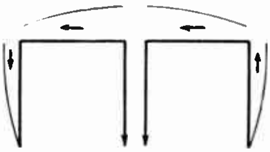

Fig. 14-20 — Folded arrangement for shortened antennas. The total length is a half-wave, not including the feeders. The horizontal part is made as long as convenient and the ends dropped down to make up the required length. The ends may be bent back on themselves like feeders to cancel radiation partially. The horizontal section should be at least a quarter wave long.

TABLE 14-II
Antenna and Feeder Lengths for Short Multiband Antennas, Center-Fed

Antenna Length (ft.)	*Feeder Length (ft.)*	*Band*	*Type of Tuning*
100	83	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

Grounded Antennas

Space restrictions often limit the size of an antenna to less than a half wavelength, particularly on 160 meters and in mobile work. In these instances, an antenna an *electrical* quarter wavelength is generally used, since it is resonant and will offer a convenient load to a line or coupling device. Quarter-wavelength antennas must be grounded at one end, so they are usually used in a vertical position, to obtain the maximum effective height.

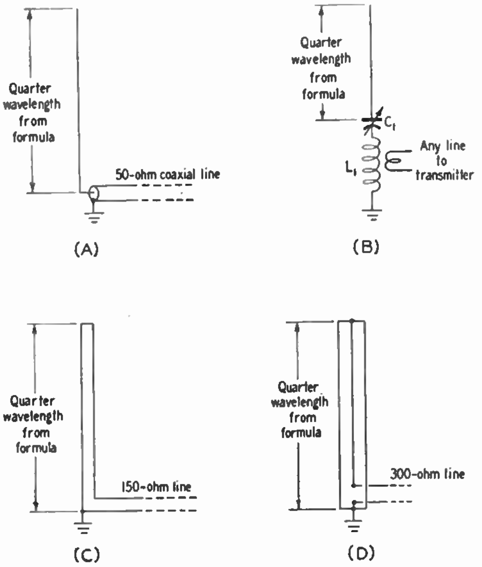

Fig. 14-21 — A quarter-wavelength antenna can be fed directly with 50-ohm coaxial line (A) with a low standing-wave ratio, or a coupling network can be used (B) that will permit a line of any impedance to be used. In (B), L_1 and C_1 should resonate to the operating frequency, and L_1 should be larger than is normally used in a plate tank circuit at the same frequency.

By using multiwire antennas, the quarter-wave vertical can be fed with (C) 150- or (D) 300-ohm line.

The impedance at the current loop of a quarter-wavelength grounded antenna is in the vicinity of 35 ohms, and thus the antenna may be fed at this point with 50-ohm coaxial cable without a serious mismatch. This and other methods of feeding quarter-wave antennas is shown in Fig. 14-21.

● 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 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. 14-22. The antenna at A uses a loading coil, L_2, to increase the electrical length of the antenna to a half wave-

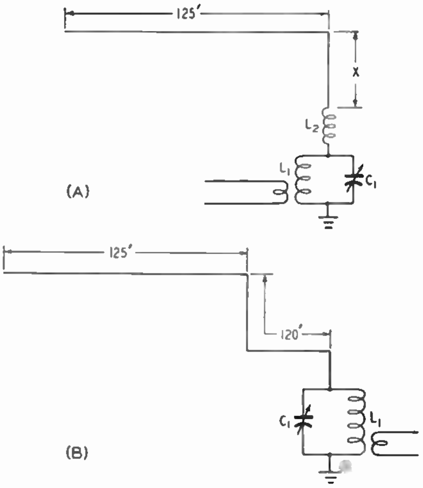

Fig. 14-22 — Bent antenna for the 160-meter band. In the system at A, the vertical portion (length X) should be made as long as possible. In either antenna system, L_1C_1 should resonate at 1900 kc., roughly. To adjust L_2 in antenna A, resonate L_1C_1 alone to the operating frequency, then connect it 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.

length, so that the antenna can be fed at its high-voltage point through the coupling circuit L_1C_1. The antenna of Fig. 14-22B 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.

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.

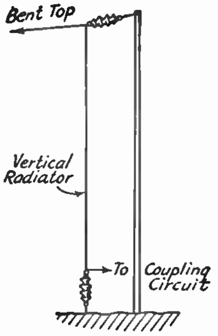

Fig. 14-23 — An arrangement for keeping the main radiating portion of the antenna vertical.

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.

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 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. A top view of the "V" antenna is shown in Fig. 14-24.

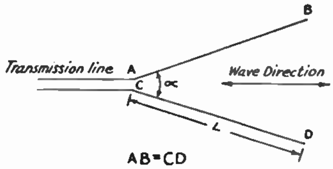

Fig. 14-24 — The basic "V" antenna, made by combining two long wires.

Fig. 14-25 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. 14-25 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

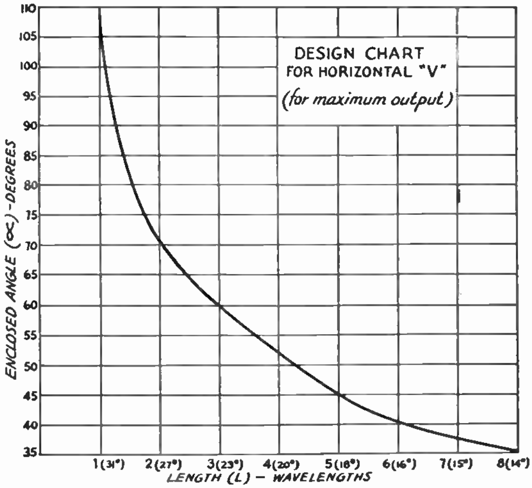

Fig. 14-25 — Design chart for horizontal "V" antennas, giving the enclosed angle between sides *vs.* the length of the wires. Values in parentheses represent approximate wave angle for height of one-half wavelength.

wires, but is not exactly twice the gain for a single long wire as given in Fig. 14-15. 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. 14-24. 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 14-G 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 in the preceding chapter.

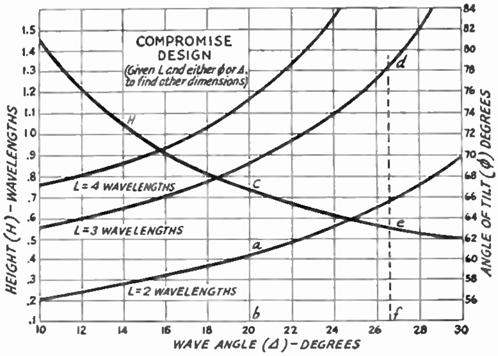

Fig. 14-27 — Compromise-method design chart for rhombic antennas of various leg lengths and wave angles. The following examples illustrate the use of the chart:

(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°.

THE RHOMBIC ANTENNA

The horizontal rhombic or "diamond" antenna is shown in Fig. 14-26. 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. 14-26, 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. 14-26. 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. 14-27 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 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.

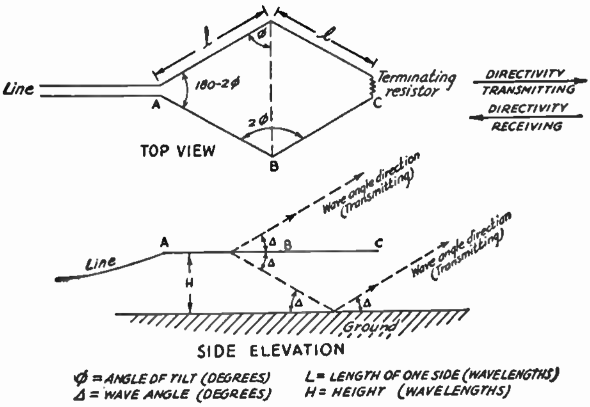

Fig. 14-26 — The horizontal rhombic or diamond antenna, terminated. Important design dimensions are indicated; details in text.

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 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 generally used with 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.**

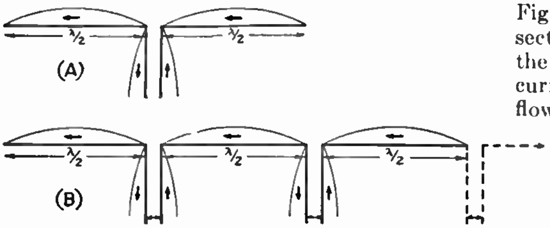

Fig. 14-28 — Collinear half-wave antennas in phase. The system at A is generally known as "two half-waves in phase." B is an extension of the system; in theory the number of elements may be carried on indefinitely, but practical considerations usually limit the elements to four.

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. 14-28. 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. 14-28B. 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 14-III. 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 14-III

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. 14-29 to form a **broadside** array, so named because 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. 14-30. Half-wave spacing generally is used, since it simplifies the problem of feeding the system when the array has more than two elements. Table 14-IV gives theoretical gain as a function of the number of elements with half-wave spacing.

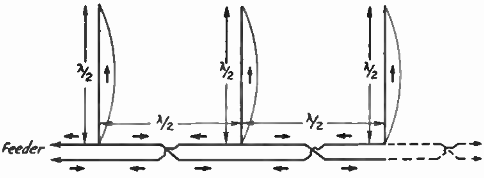

Fig. 14-29 — Broadside array using parallel half-wave elements. Arrows indicate the direction of current flow. Transposition of the feeders is necessary to bring the antenna currents in phase. Any reasonable number of elements may be used. The array is bidirectional, with maximum radiation "broadside" or perpendicular to the antenna plane (perpendicularly through this page).

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. 14-29, 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. 14-31. 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. 14-31 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. 14-31B, known as the "lazy-H" antenna, has been quite frequently used. This arrangement is shown, with the feed point indicated, in Fig. 14-32.

End-Fire Arrays

Fig. 14-33 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 plane of the antennas, as shown.

The end-fire array may be used either ver-

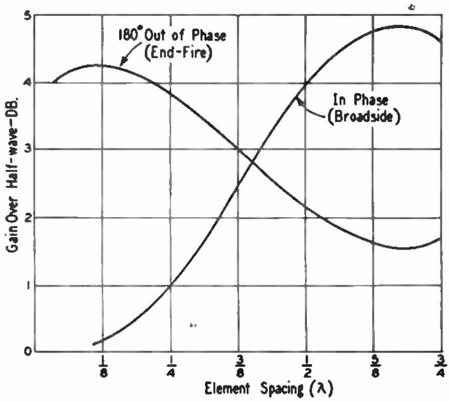

Fig. 14-30 — Gain *vs.* spacing for two parallel half-wave elements combined as either broadside or end-fire arrays.

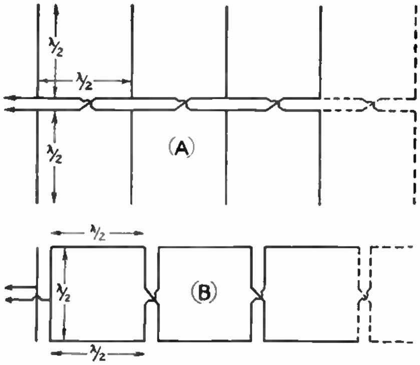

Fig. 14-31 — Combination broadside and collinear arrays. A, with vertical elements; B, with horizontal elements. Both arrays give low-angle radiation. Two or more sections may be used. The gain in db. will be equal, approximately, to the sum of the gain for one set of broadside elements (Table 14-IV) plus the gain of one set of collinear elements (Table 14-III). 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. 14-30 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.

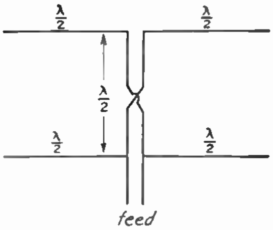

Fig. 14-32 — A four-element combination broadside-collinear array, popularly known as the "lazy-H" antenna. A closed quarter-wave stub may be used at the feed point to match into a 600-ohm transmission line, or resonant feeders may be attached at the point indicated. The gain over a half-wave antenna 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. 14-31 and 14-33 illustrate a point in connection with feeding a phased antenna system which sometimes is confusing. In Fig. 14-33, 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. 14-31B. 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. 14-29.

Adjustment of Arrays

With arrays of the types just described, using half-wave spacing between elements, it will usually suffice to make the length of each element that given by Equations **14-B** or **14-C.** The half-wave phasing lines between the parallel elements should be of open-wire construction, and their length can be calculated from:

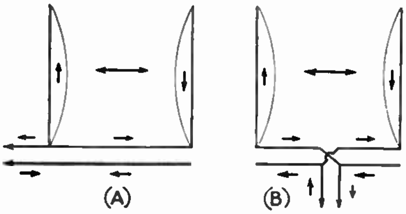

Fig. 14-33 — End-fire arrays using parallel half-wave elements. The elements are shown with half-wave spacing to illustrate feeder connections. In practice, closer spacings are desirable, as shown by Fig. 14-30. Direction of maximum radiation is shown by the large arrows.

$$\text{Length of half-wave line (feet)} = \frac{480}{\text{Freq. (Mc.)}} \qquad \textbf{(14-4)}$$

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 14-IV

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. 14-28B, 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:

$$\text{Length of quarter-wave line (feet)} = \frac{240}{\text{Freq. (Mc.)}} \quad \textbf{(14-I)}$$

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.

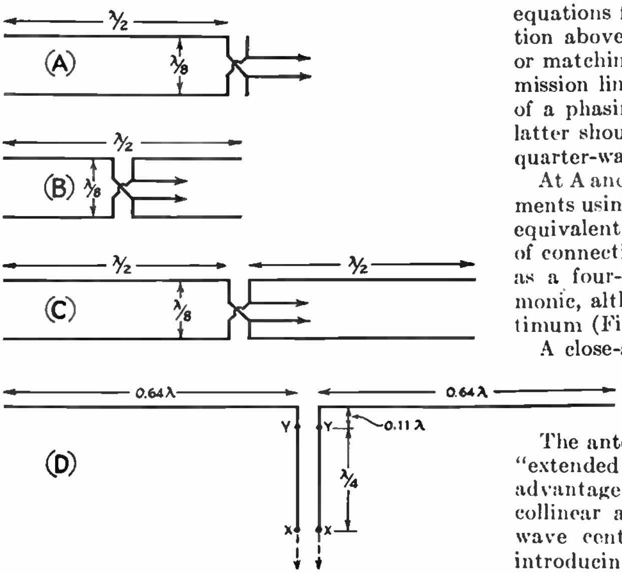

Fig. 14-34 — Simple directive-antenna systems. A is a two-element end-fire array; B is the same array with center feed, which permits use of the array on the second harmonic, where it becomes a four-element array with quarter-wave spacing. C is a four-element end-fire array with ⅛-wave spacing. D is 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 then 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 × 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. 14-28B, 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. 14-34. 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. 14-30) 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,

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.

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. 14-35, 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 reduces interference coming from the opposite direction to the signal.

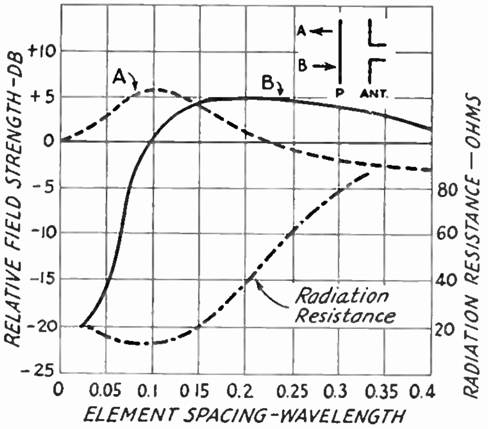

Fig. 14-35 — Gain *vs.* element spacing for an antenna and one parasitic element. The reference point, 0 db., is the field strength from a half-wave antenna alone. The greatest gain is in direction *A* at spacings of less than 0.14 wavelength, and in direction *B* at greater spacings. The front-to-back 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.

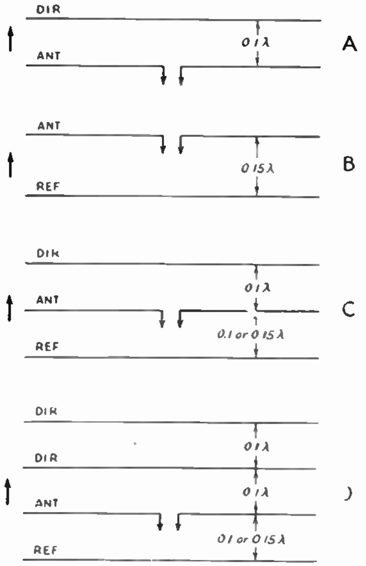

Fig. 14-36 — Half-wave antennas with parasitic elements. A, with director; B, with reflector; C, with both director and reflector; D, two directors and one reflector. Gain is approximately as shown by Fig. 14-35, in the first two cases, and depends upon the spacing and length of the parasitic element. 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.

Element Lengths

The antenna length is given approximately 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.

Simple Systems: the Rotary Beam

Four practical combinations of antenna, reflector and director elements are shown in Fig. 14-36. 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. 14-35. 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. 14-36, will give more gain and directivity than is indicated for a single reflector or director by the curves of Fig. 14-35. 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. 14-38. Resonant feeders are not recommended for lengths greater than a half-wavelength unless open-wire lines of copper-tubing conductors are used.

Three versions of the popular "T"-match are shown, for two-wire lines of Twin-Lead at A, for single coaxial line at B, and for double coaxial line at C. The match is adjusted by moving the shorting bars, keeping them equidistant from the center, until the minimum s.w.r. is obtained on the line. If the s.w.r. minimum is not 1.5 or less, the transmitter frequency should be shifted to find the frequency where the minimum s.w.r. occurs. If it is higher than the original test frequency, it indicates that the antenna element length should be increased. The parasitic element lengths taken from Fig. 14-37 should not require any adjustment, but it may be necessary to change the position of the shorting bars and the length of the antenna element once or twice before the s.w.r. at the test frequency is acceptable. The matching section may be made of the same type of conductor as the element and spaced a few inches from it. The length of the matching section will be greater with higher-impedance lines and with wider element spacing. A good starting point for a 28-Mc. wide-spaced (0.2D-0.15R) beam fed with 300-ohm Twin-Lead is 28 inches each side of center. A similar antenna and line on 14 Mc. might require about 56 inches each side.

The gamma match, shown in Fig. 14-38D, can be considered as one-half a "T"-match,

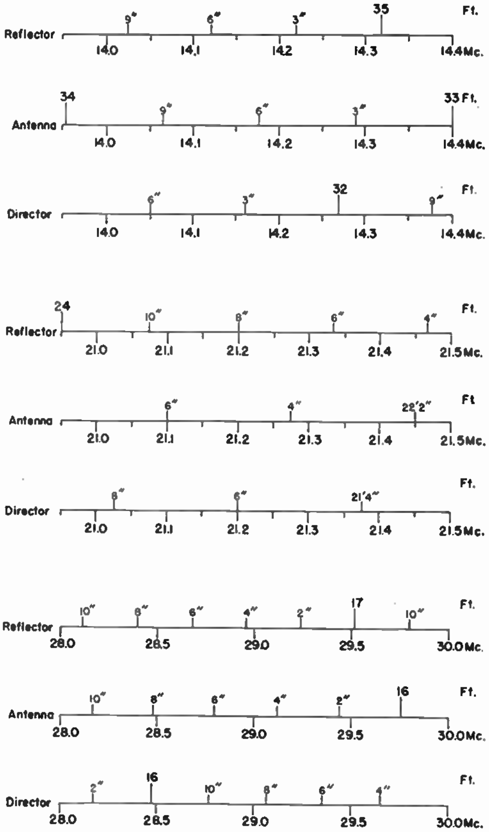

Fig. 14-37 — Director, antenna and reflector lengths for three-element beams, for element spacing of 0.1 to 0.2 wavelength. The greater spacing will result in slightly higher gain. The lengths indicated are for maximum gain — some improvement in front-to-back ratio may be obtained by adjustment of the reflector length.

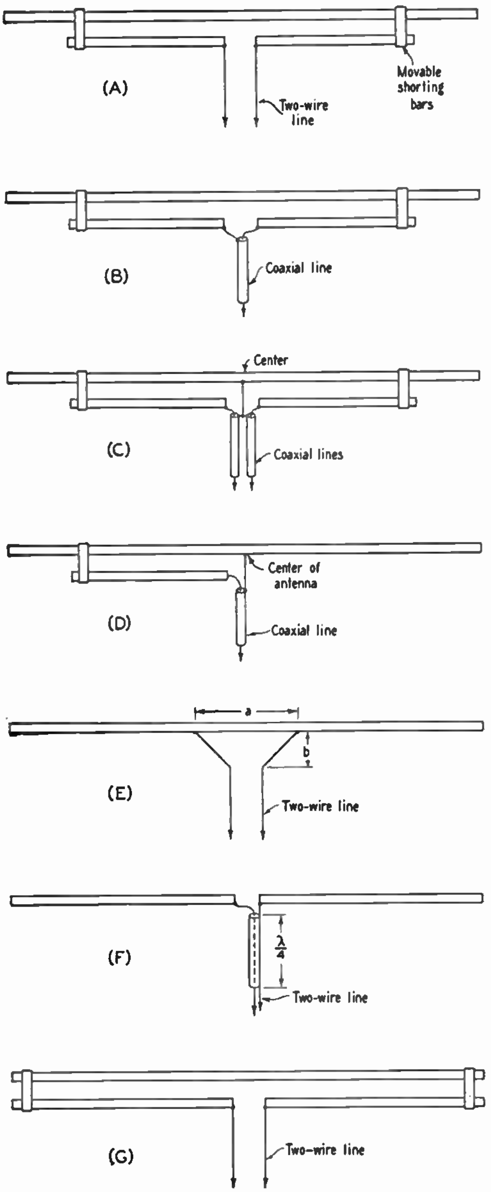

Fig. 14-38 — **Recommended methods of feeding the driven antenna element in close-spaced parasitic arrays. The parasitic elements are not shown. A, B, C, "T" match; D, "gamma" match; E, delta matching transformer; F, coaxial-line quarter-wave matching section; G, folded dipole. Adjustment details are discussed in the text.**

and the same principles hold. However, when the length of the element is changed, in an effort to minimize the s.w.r., only the side to which the movable bar is connected should be changed — the other side should remain at one-half the length obtained from Fig. 14-37. With 52-ohm coaxial line feed, the length of the matching element may run around 15 to 20 inches in a 28-Mc. beam, and twice this value in a 14-Mc. array.

The delta matching transformer shown at E 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 coaxial-line matching section at F 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 a quarter-wavelength 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

$$\textit{Length}\text{ (feet)} = \frac{246V}{f} \quad \ldots\ldots\ldots\ldots\ldots\ldots \mathbf{(14\text{–}J)}$$

where V = Velocity factor
f = Frequency in Mc.

Example: A quarter-wave transformer of RG-11/U is to be used at 28.7 Mc. From the table in Chapter Thirteen, $V = 0.66$.

$$\text{Length} = \frac{246 \times 0.66}{28.7} = 5.67 \text{ feet}$$
$$= 5 \text{ feet } 8 \text{ inches}$$

The folded-dipole antenna, Fig. 14-38G, presents a good match for the line when properly designed. Details are given in Chapter Thirteen. Different impedance step-up ratios can be obtained by varying the number of conductors or their diameter-ratio.

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 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.

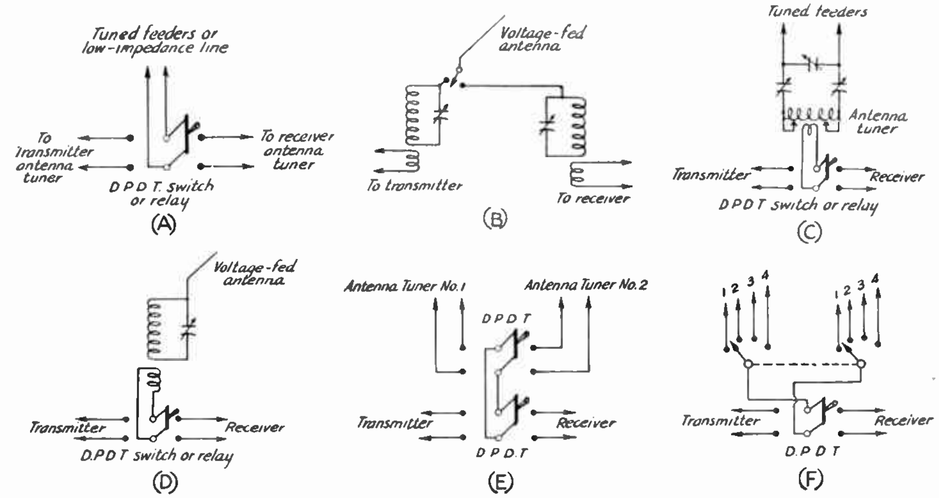

Fig. 14-39 — **Antenna-switching arrangements for various types of antennas and coupling systems. A — For tuned lines with separate antenna tuners or low-impedance lines. B — For a voltage-fed antenna. C — For a tuned line with a single antenna tuner. D — For a voltage-fed antenna with a single tuner. 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.**

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 14-III, the gain of four collinear elements is 4.5 db. with half-wave spacing; from Fig. 14-30 or Table 14-IV, the gain of two broadside elements at half-wave spacing is 4.0 db.; from Fig. 14-35, 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 it makes no difference in the final result if the array is considered as a grouping of several sets of antennas plus reflectors or as an array of antennas plus an array of reflectors. 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 location, 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 greatly reduce the directivity effects.

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. 14-39. 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 two-wire open line, 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.

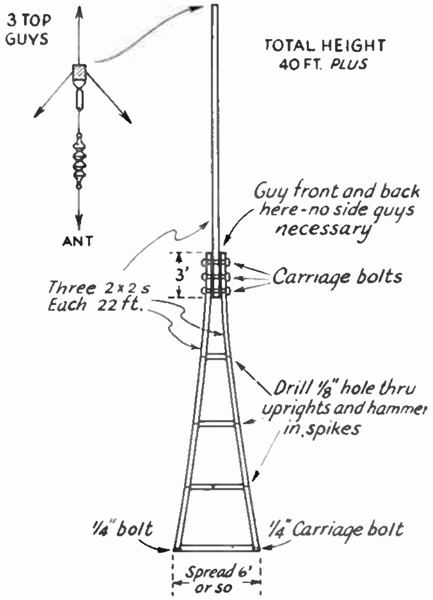

Fig. 14-40 — Details of a simple 40-foot "A"-frame mast suitable for erection in locations where space is limited.

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 from one chimney to another or from a housetop 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.

Telephone poles, if they can be purchased and installed economically, make excellent supports because they do not ordinarily require guying in heights up to 40 feet or so. Many low-cost television-antenna supports are now available, and they should not be overlooked as possible antenna aids.

● "A"-FRAME MAST

The simple and inexpensive mast shown in Fig. 14-40 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. 14-41 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 measure 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.

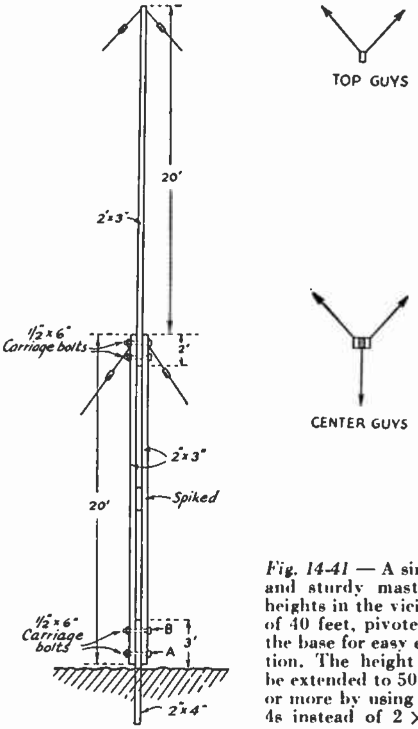

Fig. 14-41 — A simple and sturdy mast for heights in the vicinity of 40 feet, pivoted at the base for easy erection. The height can be extended to 50 feet or more by using 2 × 4s instead of 2 × 3s.

● 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. 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

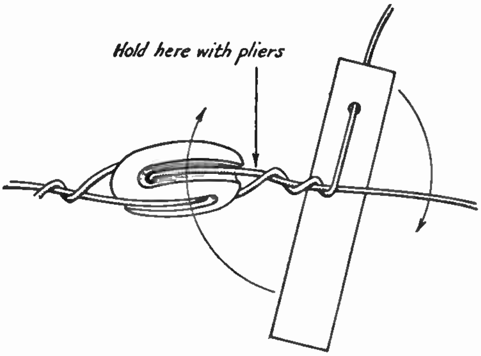

Fig. 14-42 — **Using a lever for twisting heavy guy wires.**

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. 14-42 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. 14-43.

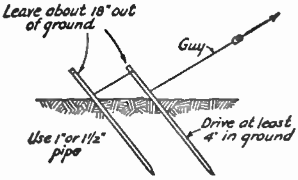

Fig. 14-43 — **Pipe guy anchors. One pipe is sufficient for small masts, but two installed as shown will provide the additional strength required for the larger poles.**

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.

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.

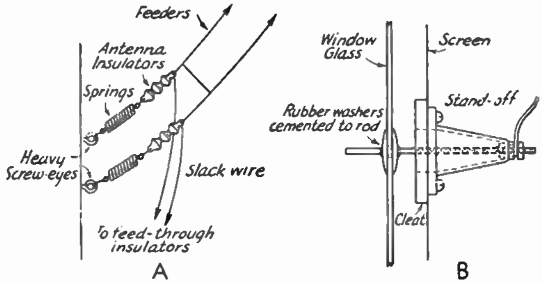

Fig. 14-44 — **A — Anchoring feeders takes the strain from feed-through insulators or window glass. B — Going through a full-length screen, a cleat is fastened to the frame of the screen on the inside. Clearance holes are cut in the cleat and also in the screen.**

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. 14-44, 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 win-

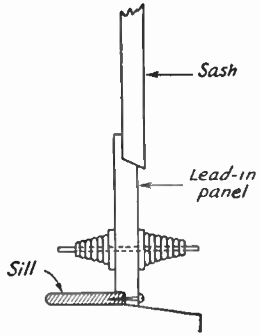

Fig. 14-45 — **An antenna lead-in panel may be placed over the top sash or under the lower sash of a window. Substituting a smaller height sash in half the window will simplify the weatherproofing problem where the sash overlap.**

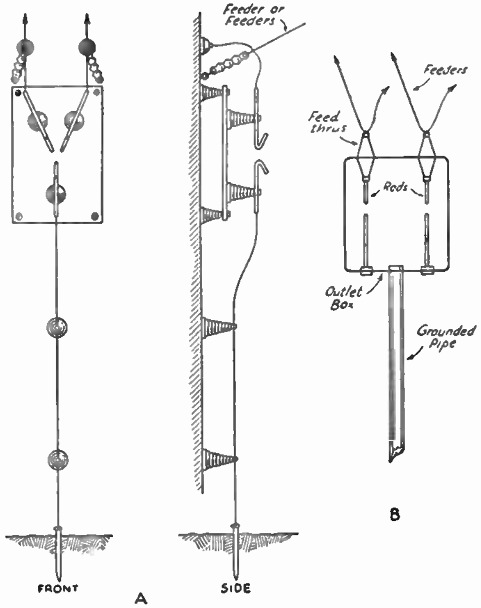

Fig. 14-46 — Low-loss lightning arresters for transmitting-antenna installations.

dow frame which provides flat surfaces for lead-in insulators. Either cement or rubber gaskets may be used to waterproof the exposed joints.

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. 14-44B 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. 14-45.

● 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. 14-46. 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 another chapter.

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.

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 should be taken to prevent rusting. Even copper-coated steel does not stand up indefinitely, since the coating usually is too thin.

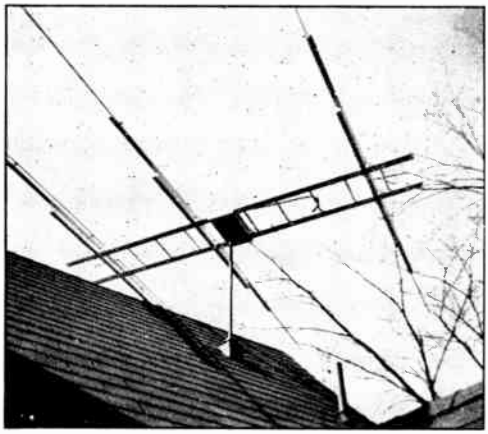

Fig. 14-47 — **A ladder-supported 3-element 28-Mc. beam. It is mounted on a pipe mast that projects through a bearing in the roof and is turned from the attic operating room. (W1MRK in August, 1946, *QST*.)**

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 or 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.

The support may be coupled to the pole by any convenient means which permits rotation or, alternatively, it may be firmly fastened to the pole and the latter rotated in bearings affixed to the side of the house.

One type of construction is shown in Fig. 14-47. It uses a section of ordinary ladder as the main support, with crosspieces to hold the tubing antenna elements

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. 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 Figs. 14-48, 14-49 and 14-50 is an example. It uses 1$\frac{3}{4}$-inch angle for the main pieces and $\frac{3}{4}$-inch angle for the other members, and the entire framework plus elements weighs only forty pounds. This simplifies considerably the problem of support.

The following aluminum pieces are required:

4 — 1-inch diameter tubing, 12 feet long. $\frac{1}{16}$-inch wall
8 — $\frac{7}{8}$-inch diameter tubing, 12 feet long, $\frac{1}{32}$-inch wall. Must fit snugly into 1-inch tubing.
2 — 1$\frac{3}{4}$-inch angle, 21 feet long
2 — $\frac{3}{4}$-inch angle, 21 feet long
4 — $\frac{3}{4}$-inch angle, 1 foot long
2 — $\frac{1}{2}$-inch diameter tubing, 6 feet long

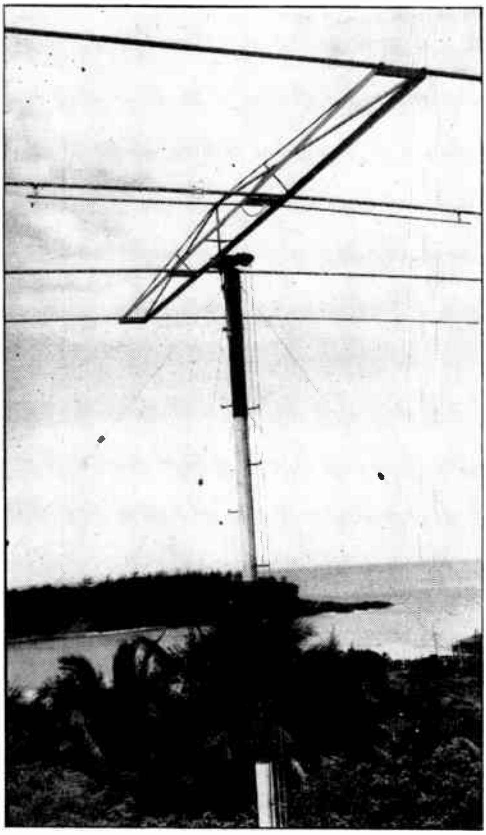

Fig. 14-48 — **A four-element 14 Mc. beam of lightweight all-metal construction. Fed by coaxial cable and hand-rotated, the antenna and boom assembly weighs only 40 pounds. (KH6IJ, Dec., 1947, *QST*.)**

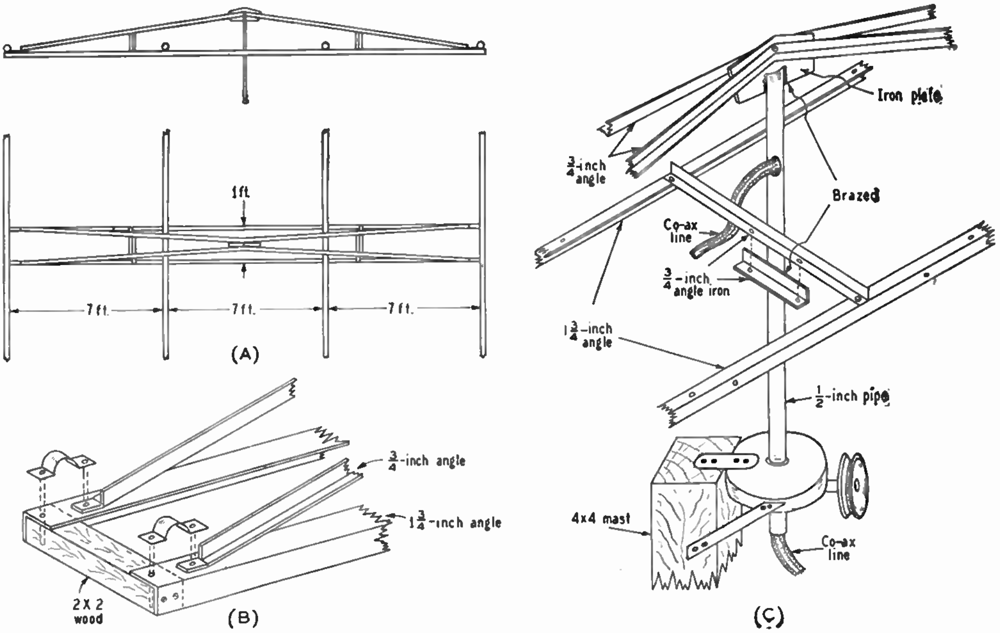

Fig. 14-49 — Details of the 4-element beam construction. The general dimensions and arrangement of the beam are given in A, the detail of the ends of the boom is shown at B, and C shows the construction of the central pivot. A discarded-forge blower gear train is used to drive the assembly.

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.

As shown in Fig. 14-49A, 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. 14-49B the spacers are recommended.

The cross braces shown in Fig. 14-50 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. 14-49C. 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

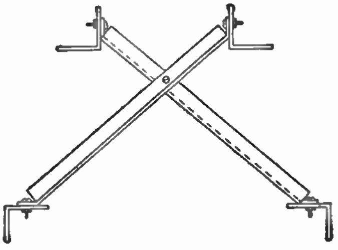

Fig. 14-50 — The boom for the 4-element beam is cross-braced at two points, about 6½ feet in from the ends.

of the house, and only the top set of guys used to provide additional support.

With all-metal construction, delta, "gamma" or "T"-match are the only practical matching methods 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. 14-51 and 14-52 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. 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 and the elements will be neat.

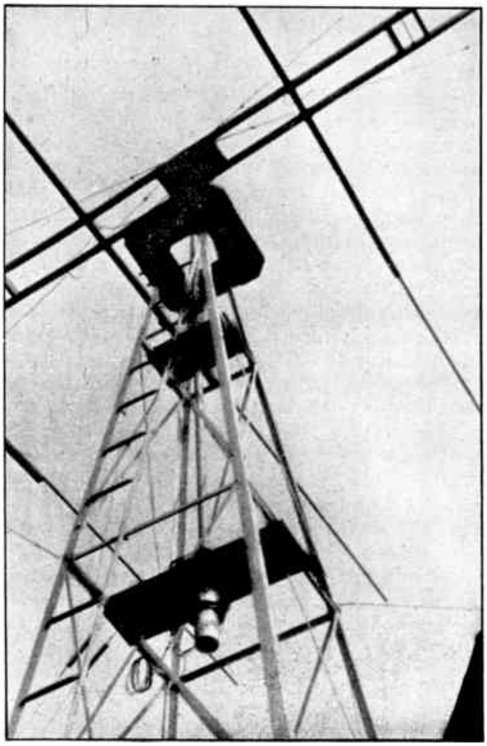

Fig. 14-51 — A wooden boom for a 4-element 14-Mc. boom can be made quite strong by judicious use of guy wires. This installation is made on a windmill tower, and the drive motor is mounted halfway down on the tower. (W6MJB, Nov., 1947, *QST*.)

The elements are 1⅜- and 1½-inch diameter duralumin tubing, supported by 1½-inch stand-off 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. 14-51 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. 14-52.

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-wavelength 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. 14-53, 14-54, 14-55, 14-56 and 14-57.

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, 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

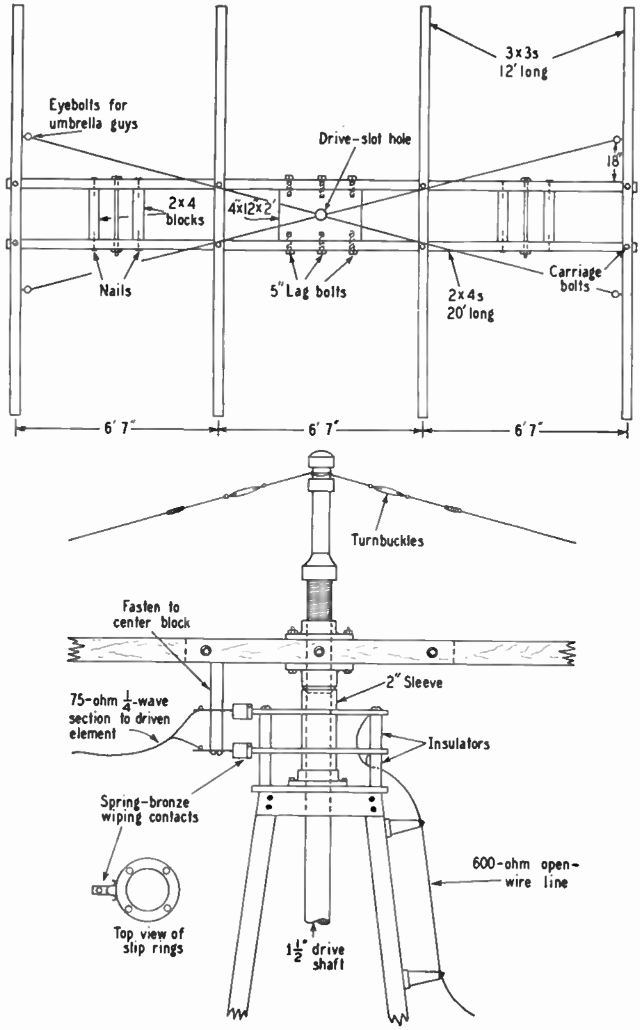

Fig. 14-52 — Details of the wooden boom, its method of support and the construction of the slip rings.

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. 14-54, while the driven element is secured in a cradle made of half sections of iron pipe welded together, as shown in Fig. 14-55. 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 impedance, 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. 14-57. 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. 14-56. 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 practicable. 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.

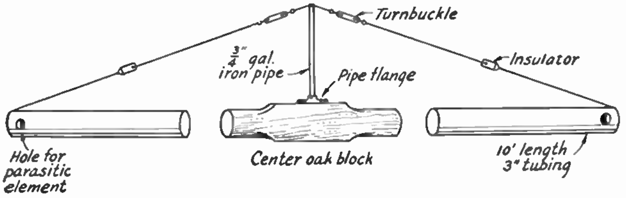

Fig. 14-53 — The boom is made of two 10-foot lengths of dural tubing slipped over a 3-foot oak block and held in place with 2-inch wood screws. Guy wires from the center add strength to the boom structure.

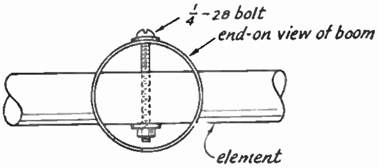

Fig. 14-54 — The center element section is held in the boom with a ¼-28 machine screw, nut and lock washer. The guy wire attaches to the head of the bolt.

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. A quarter-wave feeder system is connected to a tuned pick-up circuit whose inductance is coupled to a link. 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.

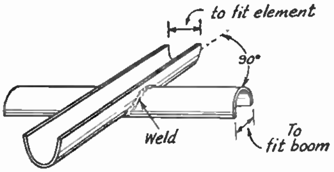

Fig. 14-55 — The clamp for the driven element is made by splitting 1-foot lengths of iron pipe and welding them as shown.

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 home-made unit must be considered. Generally speaking, lightweight units are better because they reduce the load on the mast or tower.

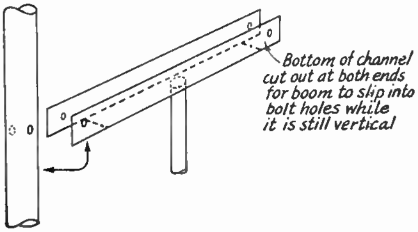

Fig. 14-56 — The mounting plate is made from a length of "U"-channel iron cut and drilled as shown. The boom is raised vertically until one set of bolt holes is in line and a bolt is slipped through. The boom is then swung into its horizontal position and the other bolt is put in place.

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

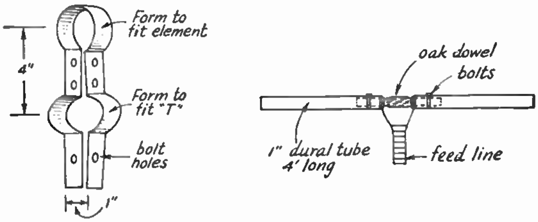

Fig. 14-57 — Details of the "T"-match assembly.

be installed in a more accessible location.

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.

CHAPTER 15

About V.H.F.

While it is possible to use the frequencies above 30 Mc. without knowing anything about wave propagation, the amateur who understands something of the means by which his signals reach distant points will be able to do a better job of it. Because much of the pleasure and satisfaction to be derived from v.h.f. work lie in making the best possible use of propagation vagaries associated with natural phenomena, a working knowledge of the basic principles of wave propagation is a most useful tool for the v.h.f. operator.

What To Expect of the V.H.F. Bands

The assignments from 50 Mc. up are superior to our lower bands in one outstanding respect: their ability to provide interference-free communication consistently within a limited service area. Lower frequencies are more subject to varying conditions that impair their effectiveness for work over a radius of 100 miles or less at least part of the time, and the heavy occupancy they support creates a continuing interference problem. Our v.h.f. bands, on the other hand, are seldom crowded, and their characteristics for local work are more stable. Because of these attributes the 50- and 144-Mc. bands, particularly, enjoy considerable popularity in areas where there are dense concentrations of population.

In addition, it has been found that there are several media by which v.h.f. signals are propagated beyond the local range, and operation on the v.h.f. bands has been taken up by many operators who must depend almost entirely on "DX" for their contacts. The latter group, particularly, will benefit from a familiarity with common propagation phenomena. The material to follow is intended to supplement the more detailed information in Chapter 4, dealing with wave propagation as it affects the world above 50 Mc.

50 to 54 Mc.

This band is borderline territory between the frequencies regularly used for long-distance communication and those normally employed for local work. Thus just about every form of wave propagation to be found throughout the radio spectrum will appear, on occasion, in the 50-Mc. region. This diversity has contributed greatly to the growing popularity of the 50-Mc. band in the amateur picture.

During the peak years of the sunspot cycle it is occasionally possible to work 50-Mc. DX of worldwide proportions, by reflection of signals from the F_2 layer. Sporadic-E skip provides opportunities for work over distances from 400 to 2500 miles or so during the early summer months, regardless of the solar cycle. Reflection from the aurora regions accounts for communication over 100 to 600-mile paths during pronounced ionospheric disturbances. The ever-changing weather pattern offers frequent opportunities for extension of the normal coverage to as much as 300 miles. This tropospheric condition develops most often during the warmer months, but may occur at any season. In the absence of any favorable propagation, the average well-equipped 50-Mc. station should be able to work regularly over a radius of 75 to 100 miles or more, depending on local terrain.

144 to 148 Mc.

Ionospheric effects are greatly reduced at 144 Mc. It is doubtful whether F_2-layer reflection ever occurs at this frequency, and sporadic-E skip is a rare phenomenon. Aurora reflection is fairly common, but the signals so reflected are generally weaker than on 50 Mc. Tropospheric effects are much more pronounced than on 50 Mc., and distances covered during favorable weather conditions are much greater than on lower bands. Air-mass boundary bending has been responsible for communication on 144 Mc. over distances in excess of 1100 miles, and 500-mile work is fairly common in the warmer months. The reliable working range under normal conditions is slightly less than on 50 Mc., when comparable equipment and antennas are used.

220 Mc. and Higher

Amateur experience on the higher bands is insufficient to provide a complete picture of what may be expected in the way of unusual propagation. There is reason to believe that tropospheric bending and duct effects become more prevalent as we go higher in frequency and that much interesting work lies in store for us when we move to the frequencies above 200 Mc. in larger numbers and with improved equipment.

Propagation Phenomena

The various known means by which v.h.f. signals may be propagated over unusual distances are discussed below.

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 in 1941, 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 instances of long-distance work on 50 Mc. by F_2-layer reflection, and as late as 1950 contacts were still being made in the more favorable areas of the world by this medium. 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.

Sporadic-E Skip

Patchy concentrations of ionization in the *E*-layer region are often responsible for re-

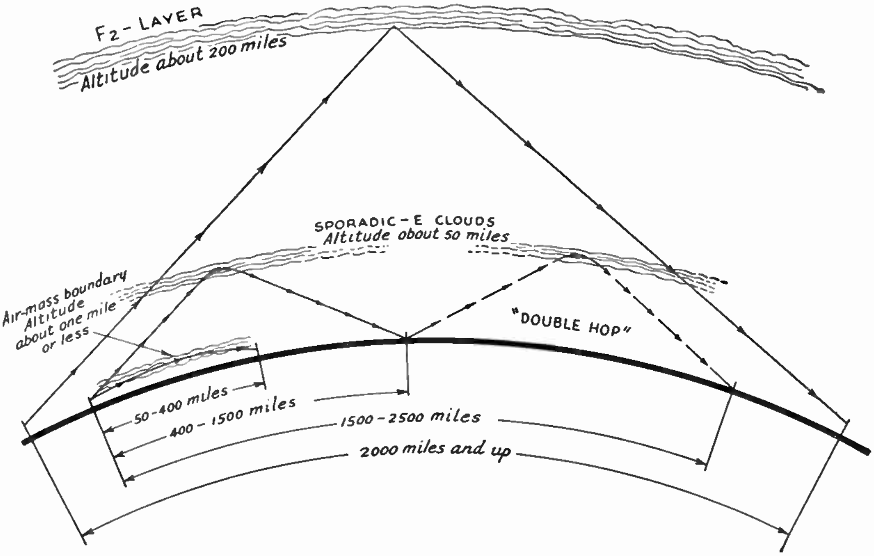

Fig. 15-1 — The principal means by which v.h.f. signals may be returned to earth. The F_2 layer, highest of the known reflecting regions of the ionosphere, is capable of reflecting 50-Mc. signals during the peak period of the 11-year solar cycle. Such communication may be world-wide in scope. Sporadic ionization of the E layer produces the familiar "short skip" contacts over medium distances at 28 and 50 Mc. On these bands it is a fairly frequent occurrence regardless of the solar cycle. It 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 communication over distances of several hundred miles, usually without a skip zone, on all v.f.h. bands.

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[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.

flection 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 upper limit of frequency for sporadic-*E* skip is not positively known, but scattered instances of 144-Mc. propagation over distances in excess of 1000 miles indicate that *E*-layer reflection, possibly aided by tropospheric effects, may be responsible.

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, v.h.f. signals may be reflected back to earth, making communication possible over distances not normally workable in the v.h.f. range. 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 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- and 144-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 800 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 many instances when it has shown up on 144 Mc. The highest frequency for aurora reflection is not yet known.

Scatter

When the maximum usable frequency for F_2-layer reflection goes above 50 Mc. it is usually possible to observe a phenomenon known to operators on 50 and 28 Mc. by a variety of terms. "Scatter" "rebound" and "reflected skip" are some of the names given to the means of propagation by which signals are returned at sharp angles from the region near the point of highest m.u.f.

The first two terms are more descriptive of what actually happens. Usually there is no skip zone, and signals so reflected may be heard over all distances up to perhaps 1000 miles. The reflection process is somewhat similar to that of aurora propagation, except that the point of reflection may be in any direction from the stations involved. Scatter signals usually show considerable audio distortion, and are subject to rapid fading.

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 v.h.f.

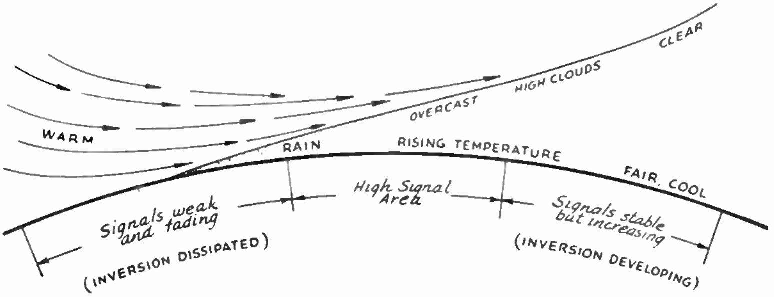

Fig. 15-2 — Illustrating a typical weather sequence, with associated variations in v.h.f. propagation. At the right is a cold air mass (fair weather, high or rising barometer, moderate summer temperatures). Approaching this from the left is a warm moist air mass, which overruns the cold air at the point 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.

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 airmass boundaries exist along the path 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 characteristics provides the medium by which v.h.f. signals cover abnormal distances. The ambitious v.h.f. enthusiast soon learns to correlate various weather manifestations with radio-propagation phenomena. By watching temperature, barometric pressure, changing cloud formations, wind direction, visibility, and other easily-observed weather signs, he is able to tell with a reasonable degree of accuracy what is in prospect on the v.h.f. bands.

The responsiveness of radio waves to varying weather conditions increases with frequency. Our 50-Mc. band is considerably more sensitive to weather variations than is the 28-Mc. band, and the 144-Mc. band may show strong signals from far beyond visual distances when the lower frequencies are relatively inactive. The maximum distance over which tropospheric propagation is frequently observed on 50 Mc. is in the neighborhood of 300 miles. On 144 Mc. distances of 500 miles are not uncommon. It is probable that this tendency continues on up through the microwave range, and that our assignments in the u.h.f. and s.h.f. portions of the frequency spectrum may someday support communication over distances far in excess of the optical range. Already 144-Mc. tropospheric communication by amateurs has passed the 1100-mile mark, and even greater distances are believed possible on this and higher frequencies.

● STATION LOCATIONS

In line with our early notions of v.h.f. wave propagation, it was once thought that only highly-elevated v.h.f. stations had any chance of working beyond a few miles. Almost all the work was done by portable stations operating from mountain tops, and only hilltop home sites were considered suitable for fixed-station work. It is still true that the fortunate amateur who lives at the top of a hill enjoys a certain advantage over his fellows on the v.h.f. bands, but high elevation is not the all-important factor it was once thought to be.

Improvements in equipment, the wide use of high-gain antenna systems, and an awareness of the opportunities afforded by weather phenomena have enabled countless v.h.f. workers to achieve excellent results from seemingly poor locations. In 50-Mc. DX work particularly, elevation has ceased to be an important factor, though it may help in extending the range of operation somewhat under normal conditions. A high elevation is somewhat more helpful on 144 Mc. and higher frequencies, particularly when no unusual propagation factors are present, as during the winter months. Other factors, such as close proximity to large bodies of water, may more than compensate for lack of elevation during the other seasons of the year, however.

Stations situated in sea-level locations along our coasts have been consistent in their ability to work long distances on 144 Mc.; weather variations provide interesting propagation effects over our Middle Western plain areas; and even the worker situated in mountainous country need not necessarily feel that he is prevented by the nature of his horizon from doing interesting work. Contacts have been made on 50 and 144 Mc. over distances in excess of 100 miles in all kinds of terrain.

The consistently-reliable nature of 50 and 144 Mc. for work over such a radius and more, regardless of weather, time or season, and the occasional opportunities these frequencies afford for exciting DX, have caused an increasing number of amateurs to migrate to the v.h.f. bands for extended-local communication, once thought possible only on the lower frequencies.

V.H.F. Receivers

Even more than in work on lower frequencies, receiver performance is all-important in the v.h.f. station. High sensitivity and good signal-to-noise ratio, necessary attributes in a receiving system for 50 Mc. and higher bands, are best attained through the use of a converter, working in conjunction with a communications receiver designed for lower frequencies. Though receivers and converters for 50, 144, and even 220 Mc. are available on the amateur market, it is possible for the v.h.f. worker to build his own with fully as good results, and at a considerable saving in cost.

In its basic principles, modern receiving equipment for these bands differs little from that employed on lower frequencies, and the same order of selectivity may be used in amateur work up to at least 220 Mc. The greatest practical selectivity should be used in v.h.f. work, as well as on the frequencies below 30 Mc., as it not only permits more stations to operate in a given band, but is an important factor in improving the signal-to-noise ratio. The effective sensitivity of a receiver having "communication" selectivity can be made considerably better than is possible with broadband systems. First on 56 Mc., more than a decade ago, then more recently on 144 Mc., and currently on 220 and 420 Mc., the change to selective superheterodyne receivers marked the beginning of real extensions of the operating range.

The superregenerative receiver, once very popular for v.h.f. work, is now used principally for portable operation, or for other applications where maximum sensitivity and selectivity are not of prime importance. It is still capable of surprising performance, for a given number of tubes and components, but its lack of selectivity, its poor signal-to-noise ratio, and its tendency to radiate a strong interfering signal rule out the superregenerator as a fixed-station receiver in areas where there is appreciable v.h.f. activity.

● R.F. AMPLIFIER DESIGN

The amount of noise generated within the receiver itself is an important factor in the effectiveness of v.h.f. receiving gear. At lower frequencies the external noise is a limiting factor, but at 50 Mc. and higher the receiver noise figure, gain and selectivity determine the ability of the system to respond to weak signals. Proper selection of r.f. amplifier tubes and appropriate circuit design aimed at low noise figure are of more importance in the v.h.f. receiver "front end" than mere gain.

Certain triode or triode-connected pentode tubes have been found superior in this respect, their superiority becoming more pronounced as we go higher in frequency. At 144 Mc., for instance, a triode r.f. stage may give substantially the same gain as a pentode, but with a much lower noise figure. With the exception of the simplest unit, the equipment described in the following pages incorporates low-noise r.f. amplifier technique.

When triodes are used as r.f. amplifiers some form of neutralization of the grid-plate capacitance is required. This can be capacitive, as is commonly used in transmitting applications, or inductive. The alternative to neutralization is the use of grounded-grid technique. Circuits for these three types of r.f. amplifier stages are given in Fig. 16-1, 16-2 and 16-3.

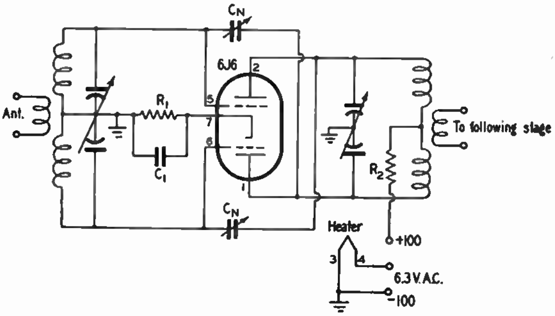

Fig. 16-1 — Schematic diagram of a pushpull r.f. amplifier for v.h.f. receiver use. This circuit is well suited to use with antenna systems fed by balanced lines. Coil and condenser sizes will be governed by the band for which the amplifier is to be used.

C_1 — 0.005-μfd. disc ceramic.
C_N — Neutralizing capacitance, about 2 μμfd. May be made from lengths of 75-ohm Twin-Lead about 1½ inches long.
R_1 — 150 ohms, ½-watt carbon.
R_2 — 1000 ohms, ½-watt carbon.

A dual triode operated as a neutralized pushpull amplifier is shown at 16-1. This arrangement is well adapted to v.h.f. preamplifier applications, or as the first stage in a converter, particularly when a balanced transmission line such as the popular 300-ohm Twin-Lead is used. It is relatively selective and may require resistive loading of the plate circuit, when used as a preamplifier. The loading effect of the following circuit may be sufficient to give the required bandwidth, when the pushpull stage is inductively coupled to the mixer.

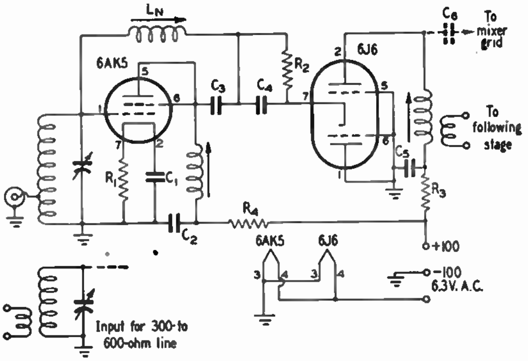

Fig. 16-2 — Circuit of the cascode r.f. amplifier. Preferred antenna coupling methods for coaxial or balanced lines are shown. The first r.f. grid coil, and the neutralizing coil, L_n, should be a high-Q design. Other coils are not critical as to Q.

C_1, C_2, C_4, C_5 — 0.005-μfd. disc ceramic.
C_3 — 50-μμfd. ceramic.
R_1, R_2 — 100 ohms, ½-watt carbon.
R_3, R_4 — 1000 ohms, ½-watt carbon.
L_n — Should resonate at signal frequency with 6AK5 grid-plate capacitance.

A two-stage triode amplifier having excellent noise figure and broadband characteristics is shown in Fig. 16-2. Commonly called the cascode, it uses a triode or triode-connected pentode followed by a triode grounded-grid stage. This circuit is extremely stable and uncritical in adjustment. At 50 Mc. and higher its over-all gain is at least equal to the best single-stage pentode amplifier and its noise figure is far lower.

Neutralization is accomplished by the coil L_n, whose value is such that it resonates at the signal frequency with the grid-plate capacitance of the tube. Its inductance is not critical; it may be omitted from the circuit without the stage going into oscillation, but neutralization results in a lower noise figure than is possible without it. Any of several v.h.f. tubes may be used in the cascode circuit, the most popular arrangement being the 6AK5-6J6 combination shown.

The neutralization process for the cascode and neutralized-triode amplifiers is somewhat similar. With the circuit operating normally the neutralizing adjustments (capacitance of C_n in Fig. 16-1; setting of slug in L_n in Fig. 16-2) can be changed until the stage stops oscillating. The middle of the range over which no oscillation occurs is approximately the proper setting. Finer adjustment can be made by disconnecting one heater lead from the r.f. amplifier tube socket and adjusting the neutralizing for *minimum* signal. A 6AK5 with one heater prong cut off may be inserted in the r.f. socket, instead of cutting the heater voltage, if desired. The best results are obtained using a noise generator, adjusting for lowest noise figure, but the two methods described above will provide a satisfactory approximation.

Grounded-grid r.f. amplifier technique is illustrated in Fig. 16-3. Here the input circuit is connected in the cathode lead, with the grid of the tube grounded, to act as a shield between cathode and plate. The grounded-grid circuit is stable and easily adjusted, and is well adapted to broadband applications. The gain per stage is low, so that two or more stages are ordinarily required. Choice of tubes is fairly limited, the best for the job being the 6J4, a triode especially designed for grounded-grid service. The 6AB4 is also suitable, and the 6J6 is used occasionally, connecting all unused elements to ground, as in Fig. 16-2. Disc-seal tubes such as the "lighthouse" and "pencil Tube" types are often used as r.f. amplifiers above 300 Mc., where ordinary miniature tubes become ineffective because of excessive lead inductance.

MIXER CIRCUITS

Triode tubes are favored for v.h.f. applications, as they are less critical as to operating conditions and the highest frequency at which they will operate satisfactorily is well above that of most pentodes. When used in mixer circuits triodes are usually quieter in operation as well.

A simple triode mixer circuit is shown in Fig. 16-4A. The grid circuit is tuned to the signal frequency, the plate circuit to the intermediate frequency. A dual-triode version is given at B. The latter is particularly suitable for use at the higher frequencies. Frequently a dual triode is used as a combination mixer-

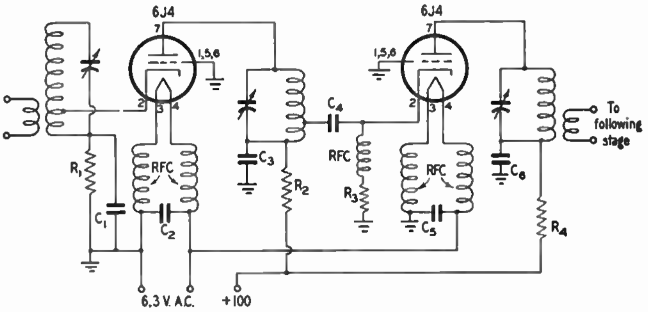

Fig. 16-3 — Grounded-grid r.f. amplifier. Position of cathode taps on coils should be adjusted for lowest noise figure.

C_1, C_2, C_3, C_5, C_6 — 0.005-μfd. disc ceramic.
C_4 — 50-μμfd. ceramic.
R_1, R_3 — 220 ohms, ½-watt carbon.
R_2, R_4 — 470 ohms, ½-watt carbon.

oscillator, using the circuits of Figs. 16-4A and 16-5A. The amount of oscillator injection is usually not critical, but in the interest of stability it should be kept as low as practical. In dual triodes having separate cathodes (7F8, 12AT7, 2C51, etc.) some external coupling may be required, but the common cathode of the 6J6 will provide sufficient injection in most cases. If the injection is more than necessary it can be reduced by dropping the oscillator plate voltage, either directly or by increasing the value of the dropping resistor, R_1.

A pentode mixer may be less subject to oscillator pulling than a triode, and it will probably require less injection voltage. If a pentode mixer is used, its plate current should be held to the lowest usable value, to reduce tube noise. This may be controlled by varying the mixer screen voltage. The principle use of pentode mixers in v.h.f. work is in the interest of simplicity of circuit layout, as in multiband converters employing bandswitching.

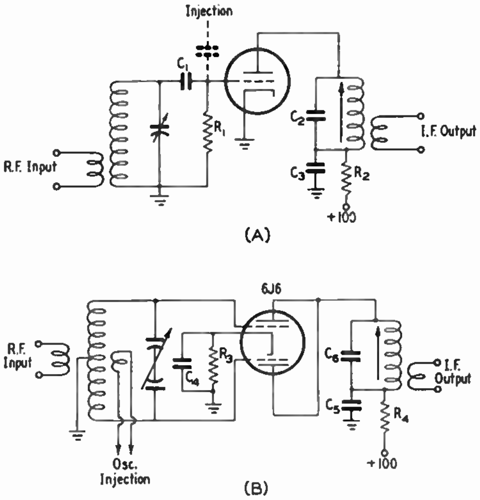

Fig. 16-4 — Two types of triode mixers suitable for v.h.f. receivers. A single-ended triode circuit is shown at A. The tube may be half of a dual triode, with the other portion used as the oscillator, or separate tubes may be used. The dual-triode version, B, is particularly useful for 144 Mc. and higher bands.

C_1 — 50 μμfd. ceramic or mica.
C_2, C_6 — 30 to 50 μμfd. ceramic or mica.
C_3, C_4, C_5 — 0.005 μfd. disc ceramic.
R_1 — 1 megohm, ½ watt.
R_2, R_4 — 1000 ohms, ½ watt.
R_3 — 150 ohms, ½ watt.

OSCILLATOR STABILITY

When a high-selectivity i.f. system is employed in v.h.f. reception, the stability of the oscillator is extremely important. Slight variations in oscillator frequency that would not be noticed when a broadband i.f. amplifier is used become intolerable when the passband is reduced to crystal-filter proportions.

One satisfactory solution to this problem is the use of a crystal-controlled oscillator, with frequency multipliers if needed, to supply the injection voltage. Such a converter usually employs one or more broadband r.f. amplifier stages, and tuning is done by varying the intermediate frequency to cover the desired frequency range.

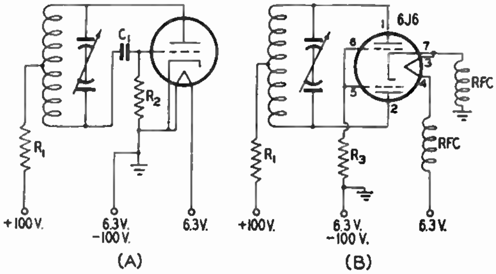

Fig. 16-5 — Recommended circuits for v.h.f. oscillators. The push-pull arrangement at B is recommended for 220 and 420 Mc., particularly.

C_1 — 50 μμfd.
R_1 — Any small carbon resistor, 1000 ohms or less.
R_2 — 10,000 ohms, ½ watt.
R_3 — 3000 to 5000 ohms, ½ watt.

When a tunable oscillator and a fixed intermediate frequency are used, special attention must be paid to the oscillator design, to be sure that it is mechanically and electrically stable. The tuning condenser should be solidly built; preferably of the double-bearing type. Split-stator condensers specifically designed for v.h.f. service, usually having ball-bearing end plates and special construction to insure short leads, are well worth their extra cost. Leads should be made with stiff wire, to reduce vibration effects. Mechanical stability of air-wound coils can be improved by tying the turns together with narrow strips of household cement at several points.

Recommended oscillator circuits for v.h.f. work are shown in Fig. 16-5. The single-ended oscillator may be used for 50 or 144 Mc. with good results. The pushpull version is recommended for higher frequencies and may also be used on the two lower bands, as well. Circuit A works well with almost any small triode, the 6AB4 or one half of a 6J6, 7F8, or 12AT7 being most commonly used. The 6J6 is well suited to pushpull applications, as shown in circuit 16-5B.

THE I.F. AMPLIFIER

Superheterodyne receivers for 50 Mc. and up should have fairly high intermediate frequencies, to reduce both oscillator pulling and image response. Approximately 10 percent of the signal frequency is commonly used, with 10.7 Mc. being set up as the standard i.f. for commercially-built f.m. receivers. This particular frequency has a disadvantage for 50-Mc. work, in that it makes the receiver subject to image response from 28-Mc. signals, if the oscillator is on the low side of the signal frequency. A spot around 7 Mc. is favored for amateur converter service, as practically all communications receivers are capable of tuning this range.

For selectivity with a reasonable number of i.f. stages, double conversion is usually employed in complete receivers for the v.h.f. range. A 7-Mc. intermediate frequency, for instance, is changed to 455 kc., by the addition of a second mixer-oscillator. This procedure is, of course, inherent in the use of a v.h.f. converter ahead of a communications receiver.

If the receiver so used is lacking in sensitivity, the over-all gain of the converter-receiver combination may be inadequate. This can be corrected by building an i.f. amplifier stage into the converter itself. Such a stage is useful even when the gain of the system is adequate without it, as the gain control can be used to permit operation of the converter with receivers of widely-different performance. If the communications receiver has an S-meter, its adjustment may be left in the position used for lower frequencies, and the converter gain set so as to make the meter read normally on v.h.f. signals.

Where reception of wide-band FM or unstable signals of modulated oscillators 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 microwave region, where stabilized transmission is extremely difficult at the current state of the art.

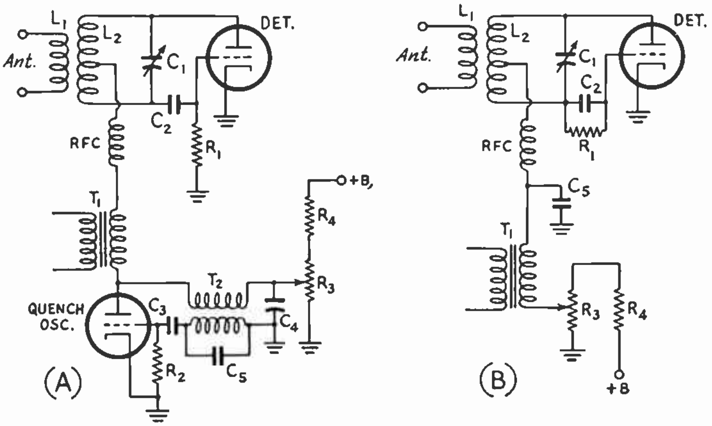

Fig. 16-6 — (A) Superregenerative detector circuit using a separate quench oscillator. (B) Self-quenched superregenerative detector circuit. L_2C_1 is tuned to the signal frequency. Typical values for other components are:

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.

THE SUPERREGENERATIVE RECEIVER

The simplest type of v.h.f. receiver is the superregenerator. It affords fair sensitivity with few tubes and elementary circuits, but its weaknesses, listed earlier, have relegated it to applications where small size and low power consumption are important considerations.

Its sensitivity results from the use of an alternating quenching voltage, usually in the range between 20 and 200 kc., to interrupt the normal oscillation of a regenerative detector. The regeneration can thus be increased far beyond the amount usable in a straight regenerative circuit. The detector itself can be made to furnish the quenching voltage, or a separate oscillator tube can be used.

When operated correctly the superregenerative receiver produces a strong hissing sound. The hiss drops or is completely eliminated when a strong signal is tuned in. Regeneration is usually controlled by varying the plate voltage in triode detectors, or the screen voltage in the case of pentodes. Typical superregenerative circuits are shown in Fig. 16-6.

The sensitivity and selectivity of the superregenerative receiver may be improved and its interference potentialities reduced by the addition of an r.f. amplifier stage ahead of the detector. This may be a broadly-tuned stage, such as any of the r.f. amplifiers shown earlier in this chapter, so that only the detector will require adjustment in tuning across the band. Such an r.f. amplifier will also make tuning of the detector less critical, as it removes the effect of antenna resonance.

The superregenerative receiver may be used effectively in mobile operation, as this type of detector combines a certain amount of a.v.c. and noise-limiting action in a single tube, and the sensitivity obtainable is greater than is possible with a superheterodyne having a comparable number of tubes. The performance of the best superregenerative receiver is not sufficiently good for the weak-signal work usually done in home stations, however, and its use on 144 Mc. for other than portable or mobile work is not recommended.

Crystal-Controlled Converters for 2, 6 and 10 Meters

The family of converters shown in Fig. 16-7 through 16-14 was designed to provide optimum performance on 28, 50 and 144 Mc. Crystal-controlled oscillators are used, to insure stability, and the triode r.f. sections provide excellent sensitivity and low noise figure. A separate "front end" for each band is plugged into a base unit containing the power supply, i.f. amplifier stage, and other parts that are not changed in shifting from one band to another.

The R. F. Circuits

The cascode circuit is used in the r.f. amplifiers of the converters for 28 and 50 Mc. A triode-connected 6AK5 with inductive neutralization works into a 6J6 grounded-grid amplifier. Circuits for the two units are similar, only the components affecting frequency being different. The functions of crystal-controlled oscillator and mixer are combined in a 6J6. The mixer plate coil is included in the plug-in unit. The schematic diagram is given in Fig. 16-8.

The 144-Mc. converter, Figs. 16-10 and 16-11, uses pushpull circuits, with a neutralized 6J6 r.f. amplifier and another 6J6 as a push-push mixer. Oscillator injection is provided by another 6J6 as crystal oscillator and multiplier. If a coaxial-line fed antenna system is used on 144 Mc. the builder may wish to use the cascode circuit on this band as well. There is little to choose from between the two circuits, except that the push-pull arrangement is better adapted to use with balanced line, and it provides about equal performance with one less tube. If the cascode circuit is used, constants for 2-meter operation will be found in Fig. 16-16.

When a fixed oscillator and variable i.f. are used, the r.f. and i.f. circuits in the converter must be made broadband, to avoid the need for readjusting them as the receiver with which the converter is used is tuned across the band. This broadbanding is accomplished in the converters for 28 and 50 Mc. by using slug-tuned plate coils in the first r.f. and mixer plate circuits. These are resonated by the circuit capacitance only, and are relatively low Q design. Coupling between the second r.f. and mixer stages employs over-coupled tuned circuits. These serve the additional purpose of providing a band-pass response, preventing interference from signals in the i.f. range. The 144-Mc. converter uses closely-coupled circuits between the r.f. and mixer stages for the same purposes. The mixer plate coil is loaded by resistor, R_4, for further broadening of the overall response.

TABLE I

Crystal And Multiplier Data For Crystal-Controlled Converters

Band (Mc.)	*Crystal (Mc.)*	*Multiplication*	*Injection Freq. (Mc.)*	*I.F. Range (Mc.)*
28 — 29.7	7.0	3rd overtone	21	7 — 8.7
50 — 54	8.6	5th overtone	43	7 — 11
144 — 148	6.8	5th overtone × 4	137	7 — 11

Crystal Oscillator Details

Crystal frequencies were selected so that the three bands would start at the same spot on the communications receiver dial, and so that the crystals would be readily obtainable from stocks. Relatively low-cost crystals are used in a regenerative triode oscillator circuit, working at an odd overtone of the crystal frequency. In the 28-Mc. unit a 7000-kc. crystal oscillates on its third overtone. Fifth-overtone operation of an 8600-kc. crystal furnishes the injection voltage in the 50-Mc. converter. A

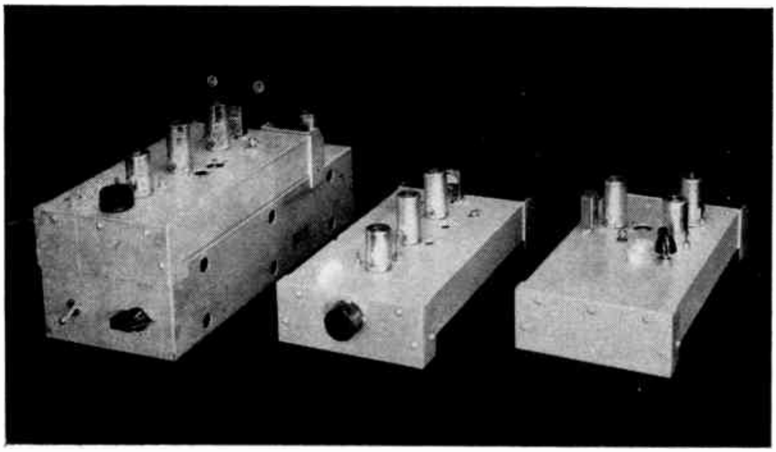

Fig. 16-7 — Crystal-controlled converters for 28, 50 and 144 Mc. At the left the 50-Mc. unit is seen mounted on the base. The latter includes an i.f. amplifier and power supply. The 28-Mc. converter (center) is similar mechanically and electrically to the 50-Mc. one. At the right is the 144-Mc plug-in unit.

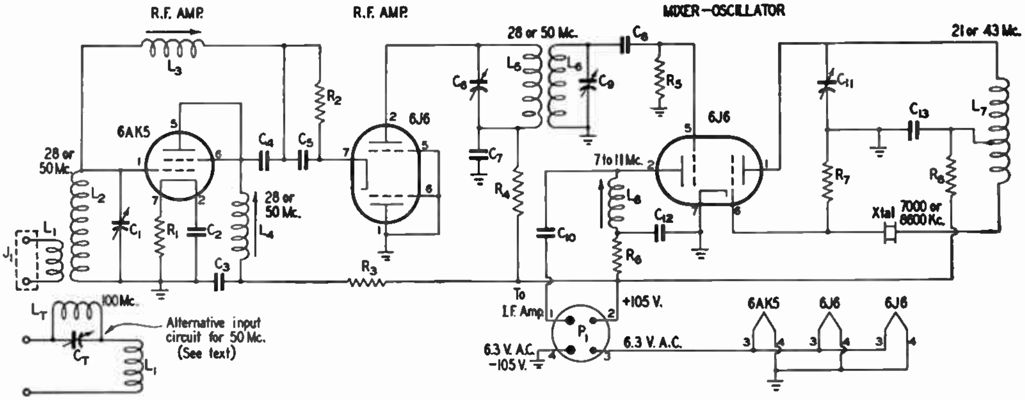

Fig. 16-8 — Schematic diagram of the crystal-controlled converters for 28 and 50 Mc. Unless otherwise indicated, parts are the same for both units.

C_1 — 15-μμfd. variable (Millen 20015).
C_2, C_3, C_7, C_{12}, C_{13} — 0.005-μfd. disc ceramic.
C_4, C_8, C_{10} — 50-μμfd. ceramic.
C_5 — 500-μμfd. ceramic.
C_6, C_9 — 5–20-μμfd. ceramic trimmer.
C_{11} — *50 Mc.:* 50-μμfd. air trimmer (Millen 26050). *28 Mc.:* 75-μμfd. air trimmer (Millen 26075).
R_1, R_2 — 100 ohms, ½ watt.
R_3, R_4, R_6, R_8 — 1000 ohms, ½ watt.
R_5 — 0.68 megohm, ½ watt.
R_7 — 3300 ohms, 1 watt.
L_1 — 4 turns No. 28 e. between turns of L_2 at cold end.
L_2 — *50 Mc.:* 10 turns No. 20 tinned, ½-inch diam., ⅝ inch long (B & W Miniductor 3003). *28 Mc.:* 14 turns No. 20 tinned, ⅝-inch diam., ⅞-inch long (B & W Miniductor 3007).
L_3 — *50 Mc.:* 25 turns No. 32 e., close-wound on CTC LSM form (¼-inch diam., slug-tuned). *28 Mc.:* CTC LS3 10-Mc. coil, slug-tuned.
L_4 — *50 Mc.:* Slug-tuned plate coil CTC LS3 30 Mc. *28 Mc.:* CTC LSM 10-Mc. coil with 4 turns removed, slug-tuned.
L_5, L_6 — *50 Mc.:* 8 turns No. 18 tinned, ⅝-inch diam., 1 inch long (B & W Miniductor 3006), ¼ inch space between cold ends. *28 Mc.:* 9 turns No. 24 tinned, ½-inch diam., 9/32 inch long (B & W Miniductor 3004), 3/16 inch space between cold ends.
L_7 — *50 Mc.:* 10 turns No. 20 tinned, tapped 3⅛ turns from crystal end (B & W Miniductor 3003), ½-inch diam., ⅝ inch long. *28 Mc.:* 10 turns No. 20 tinned, ⅝-inch diam., ⅝ inch long, tapped 3⅛ turns from crystal end (B & W Miniductor 3007).
L_T, C_T — F.m. trap. 7 turns No. 20 tinned, ½-inch diam., ⅜ inch long (B & W Miniductor 3003), tuned with 5–20-μμfd. ceramic trimmer.
J_1 — Crystal socket for antenna terminals.
P_1 — 4-prong male plug.

6800-kc. crystal oscillates on its fifth overtone in the 144-Mc. converter, multiplying by four in the second 6J6 triode section. Table 16-I gives the complete information.

Operation of crystals in this way results in a frequency that may not be an exact multiple of the frequency marked on the crystal holder; hence the term, "overtone." It is close enough for ordinary dial calibration purposes, however. Harmonic-type crystals of the proper frequency could be obtained on order, but the cost would be materially higher. Conventional operation of lower-frequency crystals, making up the multiplication with additional stages, is not recommended, because of the difficulty in avoiding birdies from crystal harmonics. In the overtone circuit, no frequency lower than the overtone at which the crystal oscillates is heard.

Layout

The units are built on aluminum chassis of stock sizes. The base is 3 by 5 by 13 inches (ICA 29003), and the r.f. units are 1½ by 5 by 9½ inches (ICA 29001). The only metal work required is the making of small aluminum guide plates for the front and rear of the converter chassis, and the mounting bracket for

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Fig. 16-9 — Bottom view of the 28-Mc. plug-in unit. At the left is the tuned input circuit, followed by the 6AK5 r.f. stage, with its slug-tuned plate and neutralizing windings. At the middle of the chassis is the 6J6 grounded-grid stage, with its bandpass coupling to the mixer grid. Oscillator components are at the upper right. Parts arrangement in the 50-Mc. converter is similar.

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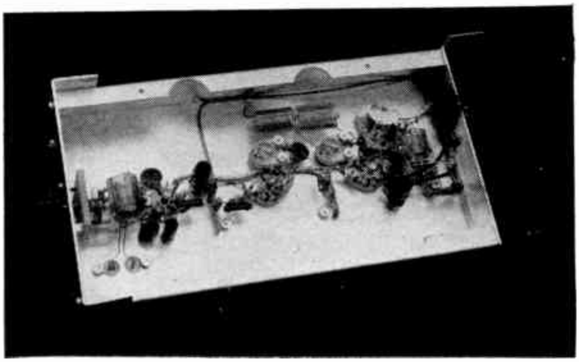

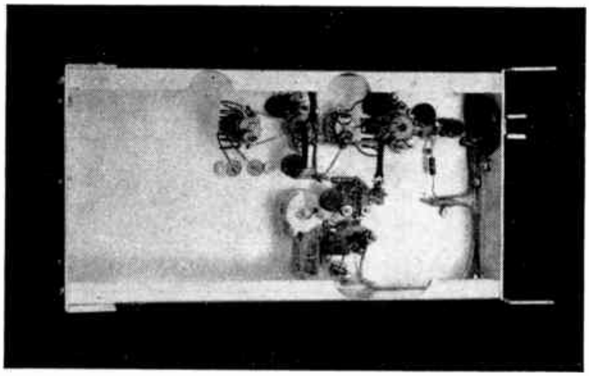

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Fig. 16-10 — Bottom view of the 144-Mc. converter. Across the top of the photo, *left to right*, are the input circuit, the push-pull r.f. stage, the push-push mixer, and its slug-tuned plate circuit. Oscillator and multiplier components are at the bottom of the picture.

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the inter-connecting socket at the rear of the base unit. Ventilation holes are cut in the sides of the base unit, and two 1¼-inch holes are cut in the top surface of this chassis to provide greater clearance around the major coils of the r.f. assemblies, when they are in the operating position. The placing of the power supply and i.f. amplifier components on the base unit is not critical, though the arrangement shown in the photographs works out nicely from a mechanical standpoint. Chief consideration here is to avoid mounting parts on the outside walls of the units, thereby preserving to the fullest degree the deep-but-narrow form factor. This shape takes up a minimum of high-priority space on the operating table.

Care should be used in mounting the socket and plug on the base unit and converters, respectively, in order that they may line up exactly. When the job is properly done it is merely necessary to place the converter unit on the base, with the front edge tilted upward slightly, slide the plug into the socket, and then drop the converter in place. The converter assemblies should be kept free of parts in the portion

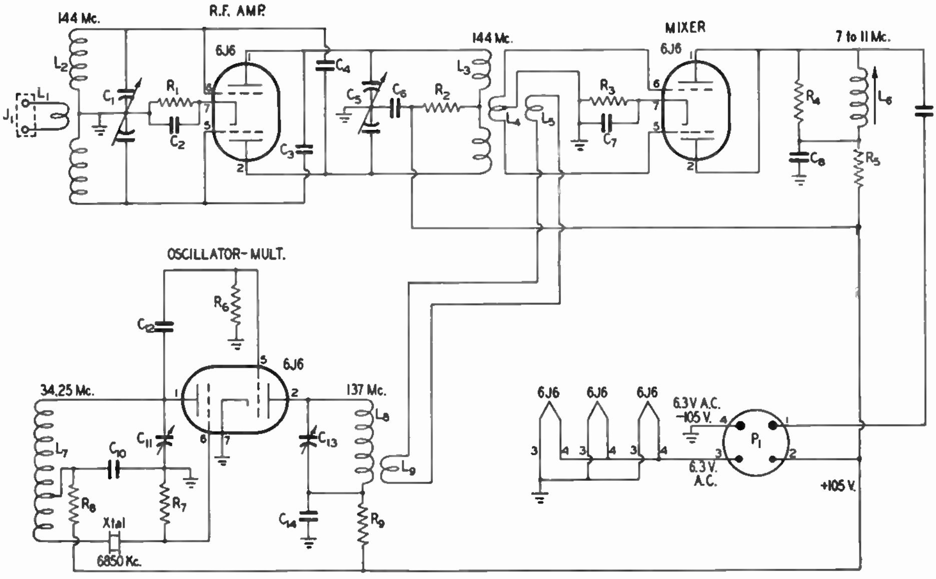

Fig. 16-11 — Wiring diagram of the 144-Mc. crystal-controlled converter.

C_1, C_5 — 5.3-μμfd.-per-section butterfly (Johnson 5MB11).
C_2, C_6, C_7, C_8, C_{10}, C_{14} — 0.005-μfd. disc ceramic.
C_3, C_4 — 75-ohm Twin-Lead neut. capacitors (see text).
C_9 — 50-μμfd. ceramic.
C_{11} — 50-μμfd. air trimmer (Millen 26050).
C_{12} — 100-μμfd. ceramic.
C_{13} — 5-20-μμfd. ceramic trimmer.
R_1, R_3 — 150 ohms, ½ watt.
R_2, R_5, R_7, R_9 — 1000 ohms, ½ watt.
R_4 — 2200 ohms, ½ watt.
R_6 — 0.22 megohm, ½ watt.
R_8 — 3300 ohms, 1 watt.
L_1 — 4 turns, No. 18 enam., 5/16-inch diam., ¼ inch long.
L_2, L_3 — 6 turns No. 18 enam., 3 turns each side of center tap, with ⅜-inch spacing between sections. ⅜-inch diam. Adjust turn spacing as needed.
L_4 — 5 turns No. 18 enam., ⅜-inch diam., close-wound and center-tapped.
L_5, L_9 — 1 turn hook-up wire wound around L_5 and L_9: 75-ohm Twin-Lead used to connect between the two coils.
L_6 — Slug-tuned plate coil (CTC LS3 5-Mc. coil with 20 turns removed).
L_7 — 11 turns No. 20 tinned, ½-inch diam., 11/16 inch long, tapped 4 turns from crystal end of coil (B & W 3003).
L_8 — 3 turns No. 18 tinned. ½-inch diam., ⅜ inch long (B & W 3002).
J_1 — Crystal socket for antenna terminal.
P_1 — 4-prong male socket.

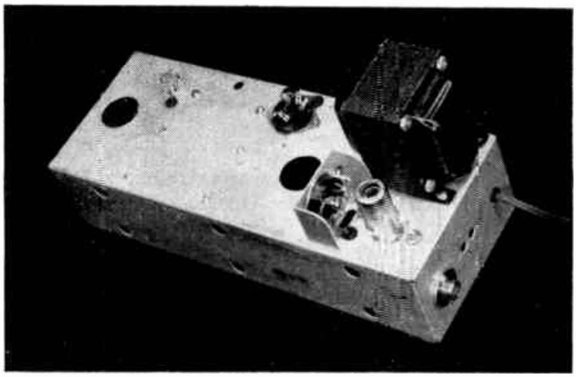

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Fig. 16-12 — Base unit, with converter removed, showing the plug-in fitting for the mixer output and power connections. The 6BA6 i.f. amplifier stage is at the lower right.

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that is over the rectifier tube socket, in order that no components be damaged in the plugging-in operation.

Looking at the converters for 28 and 50 Mc. from the front we see the tuning condenser for the r.f. input circuit, followed by the 6AK5 and 6J6 r.f. stages and the 6J6 mixer-oscillator, in that order. The 6AK5 plate coil, the neutralizing coil, and the mixer plate coil are slug-tuned resonating with the circuit capacitances only. Condenser-tuned circuits are used in the r.f. input, second r.f. plate, and mixer grid circuits. The difference in position of the r.f. tuning condenser, C_1, in the two converters is the result of an improved parts arrangement used in the 28-Mc. job. Mounting of this condenser on the front wall of the converter chassis is recommended for both units.

Note the alternative input circuit for the 50-Mc. converter, shown in Fig. 16-8. This includes a 100-Mc. trap for elimination of f.m. interference. If the converter is to be used in a location near to f.m. broadcast stations this trap is necessary to prevent the second harmonic of the injection frequency from beating with the f.m. signals and producing spurious responses in the 50-Mc. band.

In the 2-meter converter the r.f. and mixer tubes are in line at the right side of the chassis, as viewed from the front, with the oscillator-multiplier at the left. This layout makes for symmetrical arrangement of the push-pull circuits. All the r.f. coils are self-supporting, so that their length and coupling can be adjusted readily. Link coupling of the injection voltage is accomplished with single-turn coils around the multiplier-plate and mixer-grid windings, connected by a short length of 75-ohm Twin-Lead.

Adjustment and Operation

Work on the r.f. sections is made easier if a patch cord is made up so that the r.f. units can be removed from the base and kept in operating condition. The only critical portion of the adjustment procedure is that involved in getting the crystal oscillator to work properly, and on the right overtone. The important factor here is the amount of regeneration,

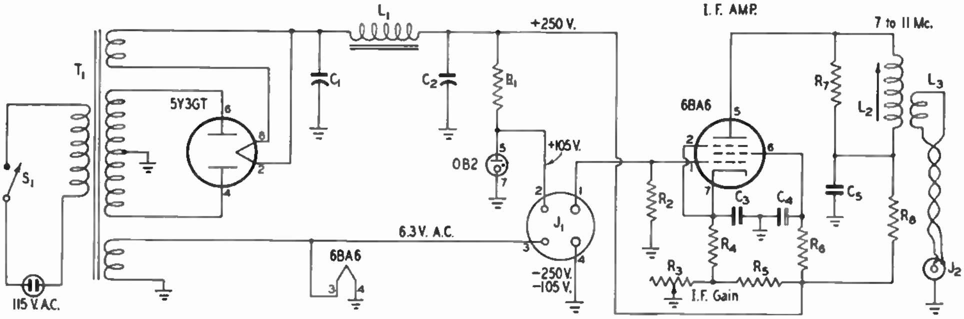

Fig. 16-13 — Wiring diagram of the power supply and i.f. amplifier unit for use with the crystal-controlled converters.

C_1, C_2 — 10-μfd. 450-volt electrolytic.
C_3, C_4, C_5 — 0.005-μfd. disc ceramic.
R_1 — 2500 ohms, 10 watts.
R_2 — 1 megohm, ½ watt.
R_3 — 10,000-ohm wire-wound potentiometer.
R_4 — 68 ohms, ½ watt.
R_5 — 56,000 ohms, 2 watts.
R_6 — 39,000 ohms, 1 watt.
R_7 — 2200 ohms, ½ watt.
R_8 — 1000 ohms, ½ watt.
L_1 — 10-hy. 50-ma. filter choke.
L_2 — Slug-tuned plate coil (CTC LS3 5 Mc. with 10 turns removed).
L_3 — 15 turns No. 32 enam., scramble-wound at bottom end of L_2.
J_1 — 4-prong female plug.
J_2 — Coaxial-cable jack.
S_1 — S.p.s.t. toggle switch.
T_1 — Power transformer, 275 v. each side c.t. at 50 ma.; 6.3 v. at 2.5 amp.; 5 v. at 2 amp. (Thordarson T-22R30).

controlled by the position of the tap on the oscillator coil, L_7. The process is the same for all three converters, but the tap position may be somewhat more critical in the 50- and 144-Mc. units, as a higher-order overtone is used.

The proper position for the tap is that at which oscillation takes place only at the third or fifth overtone, as the converter requires. If the tap is too high on the coil oscillation will be on random frequencies, determined by the setting of C_{11}, rather than controlled by the crystal. If the tap is too low on the coil no oscillation at all will develop. The L/C ratio in the tuned circuit is also fairly critical, for best operation, but if the values given in the parts lists are followed no trouble should be encountered on this score.

To check operation of the oscillator insert a meter in series with R_8, apply plate voltage, and rotate C_{11} until a sharp dip in plate current occurs, indicating oscillation. There may be a tendency to self-oscillation at the minimum-capacity end of the tuning range, but this may be disregarded if it disappears quickly as the condenser is turned toward maximum capacity. Crystal oscillation should occur somewhere between half and maximum capacity. It is helpful if a receiver is available for listening on the frequency of oscillation (indicated over L_7 in the diagrams) to see whether or not the crystal is controlling the frequency. If the frequency changes markedly or if pronounced hand-capacity effects are present, move the tap toward the low end of L_7 by one turn and try again. A fraction of a turn change may be necessary in some instances, to achieve crystal control without random oscillation. It is also possible that the wrong overtone may develop. With incorrect values of inductance and capacity this type of circuit may produce oscillation on any odd overtone, so a wavemeter or receiver check should be made to be certain that the proper injection frequency is being used.

Next a rough alignment of the r.f. and i.f. circuits should be made. This can be done on noise, with the receiver set at the approximate midpoint of the frequency range to be tuned, or if one has a signal generator the process is made easier. This need be nothing more than the crystal oscillator in the transmitter, using the proper harmonic.

Neutralizing is next in order. This should be done following the procedure outlined in the section on r.f. amplifier design earlier in this chapter.

Final adjustment of the converters may now be made. Peak all circuits in the 10- and 6-meter converters at one end of the band, then move the receiver to the other end of the band and repeak either the mixer or i.f. amplifier plate winding for maximum response. Receiver noise is satisfactory for this test. If the response is not sufficiently broad, correction can be made with the bandpass circuits in the second r.f. plate and mixer grid circuits, stagger tuning these and the i.f. coils until reasonably flat response is attained. All this is best done with a 300-ohm resistor connected across the antenna terminals, to eliminate antenna resonance effects. If the response is flat with this set-up, variation in noise over the band with the antenna on may be disregarded, since it is a function of the antenna itself. Absolutely flat response is not important, for the overall gain of the system can be adjusted by means of the i.f. gain control. It should be set so that, with the antenna connected, the normal noise level just starts to read on the meter. Turning the gain beyond the point at which noise becomes a limiting factor effects no improvement in signal readability.

The flatness of response in all converters can be varied by adjusting the r.f.-mixer coupling. In the 2-meter unit the coupling between L_3 and L_4 should be increased to the point where it is unnecessary to change the setting of C_5 to cover the entire band. There will be a slight amount of repeaking of C_1 necessary in all converters, though it should not make more than about one S-unit difference from one end of the band to the other, and it will have a negligible effect on the noise figure.

The converters are now ready for use, but some work on the receiver may be needed. A few communications receivers radiate harmonics of the high-frequency oscillator frequency, and these will show up as birdies throughout the v.h.f. range. The cure is similar to that employed in treating transmitters for TVI.

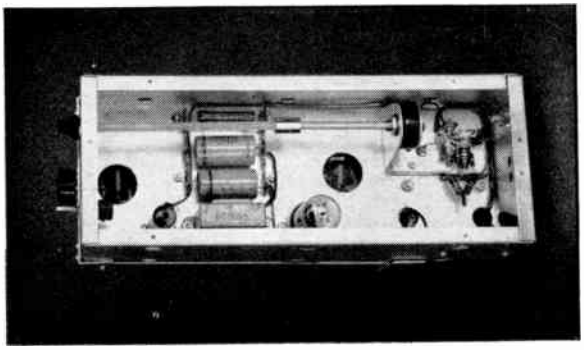

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Fig. 16-14 — **Under-chassis view of the base unit, showing the power supply and i.f. amplifier components. The circular cutouts provide additional clearance around the tuned circuits in the plug-in unit.**

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A Low-Noise Converter for 144 Mc.

The 2-meter converter shown in Figs. 16-15 to 16-18 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 cascode circuit treated earlier in this chapter. 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, if the i.f. circuits are suitably altered.

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, as may be seen in the rear view, Fig. 16-18.

The oscillator condenser 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

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Fig. 16-15 — 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.

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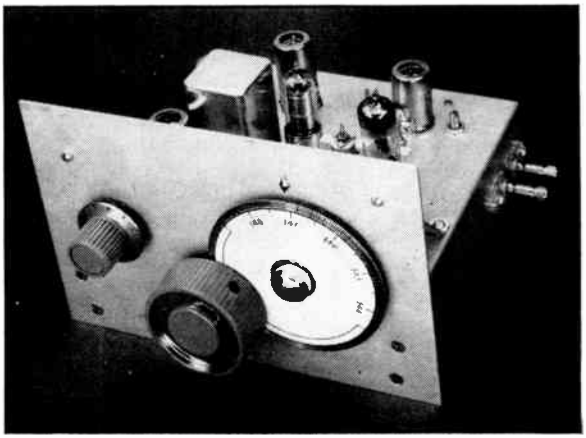

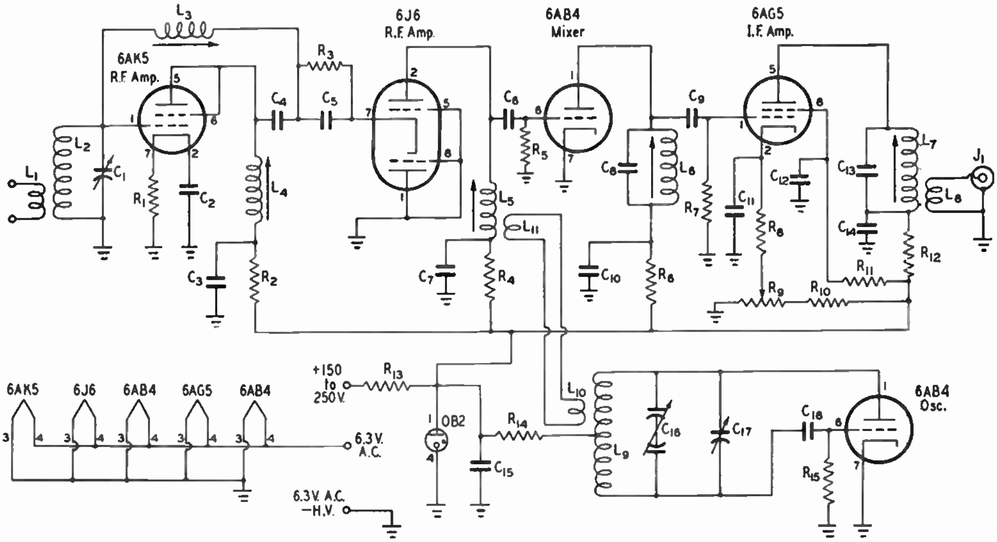

Fig. 16-16 — Schematic diagram of the 2-meter cascode converter.

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. 16-17 and 16-18. 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. 16-17.

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

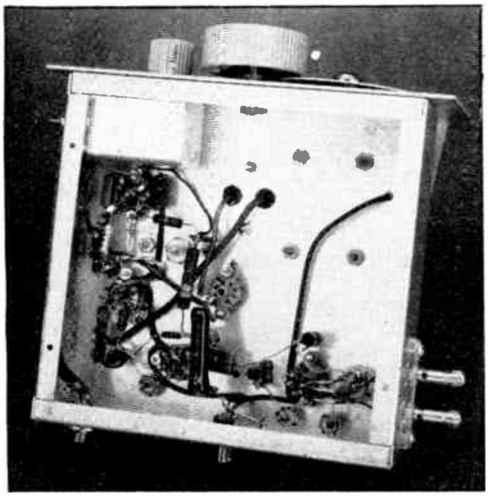

Fig. 16-17 — Bottom view of the 2-meter converter.

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.

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Fig. 16-18 — 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.

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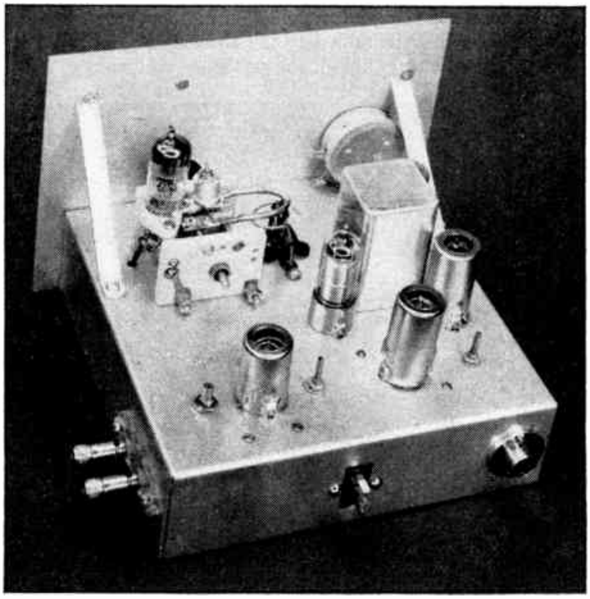

A Simple Converter for 50 and 144 Mc.

Though the more complex equipment already described is typical of the gear that must be used in order to attain top performance on the v.h.f. bands, it is possible to start with simpler devices and still do a good job. The converter shown in Figs. 16-19 through 16-22 provides the best performance that can be expected from simple equipment. It was not built to be the simplest possible receiving device; rather, it was designed to provide good results with a minimum of complication and cost.

It uses a dual triode, 6J6, as a combined mixer-oscillator, followed by a 6AK5 i.f. amplifier. The latter is necessary; do not try to do without it. The output of a triode mixer is too low to give adequate gain for most receivers. The i.f. amplifier stage makes the converter usable with even the simplest receivers, and provides a convenient means of controlling the overall gain of the system. Plug-in coils with a miniature-tube type of base provide the means of changing bands.

Mechanical Details

Though it could be built in a much smaller space, the converter uses a 3 by 5 by 10-inch chassis, allowing plenty of room for the work that must be done underside. The main tuning condenser is a split-stator variable made from a double-bearing double-spaced 35-$\mu\mu$fd. type. The stator bars are sawed at the middle and each section is reduced to four stator and three rotor plates. This unit is mounted under the chassis, as close to the top plate as possible, to make room for the vernier dial on the front panel. To provide shielding without the necessity for individual shield cans, the mixer and i.f. plate coils, L_4 and L_5, are mounted under the chassis. Normally this will provide all the shielding necessary for the i.f. circuits. If trouble is experienced with signals on the intermediate frequency a bottom plate may be added to the chassis.

A smooth-running dial on the oscillator tuning is a necessity in a v.h.f. converter. The friction-drive dial used (National Type K) is relatively inexpensive, and if a large knob is substituted for the small one with which the dial is equipped, it provides a very satisfactory tuning rate.

The circuit is so simple that no trouble should be experienced if the general parts arrangement is followed. Look over the photographs closely before starting to lay out the chassis for drilling. In the rear view, Fig. 16-20, the oscillator coil, the 6J6 tube, and the mixer grid coil, L_1-L_2, appear in that order, from left to right, close to the panel. The 6AK5 tube is nearer the back, with the slug adjustment screws of the mixer plate coil, L_4, and the i.f. plate coils, L_5-L_6, at the left and right, respectively. Looking at the bottom view, Fig. 16-19, the oscillator tuning condenser, C_5, is at the left, with its parallel trimmer, C_4, mounted directly on the stator bars, on the left side. Note that the oscillator coil socket is mounted directly under C_5, on the same center line, allowing connections from C_5 to the socket to be made with the shortest possible leads.

The only critical job in the construction or adjustment procedure is involved in getting the inductance of the oscillator plug-in coils, L_3, to the correct value. There being only one parallel trimmer for the oscillator (C_4) the coils must be made and adjusted carefully in order to have the desired bandspread on both ranges.

Considerable care must be used in the placement of the oscillator and mixer components, so that all leads will be very short; otherwise it will not be possible to resonate these circuits at 148 Mc. The 6J6 socket is at the right of C_5 in the bottom view, and the mixer grid circuit components appear just to the right of the middle. The i.f. amplifier gain control, R_7, is at the right. The 300-ohm line from the crystal-socket antenna terminal, J_1, may be seen at the far right. The mixer plate coil, the i.f.

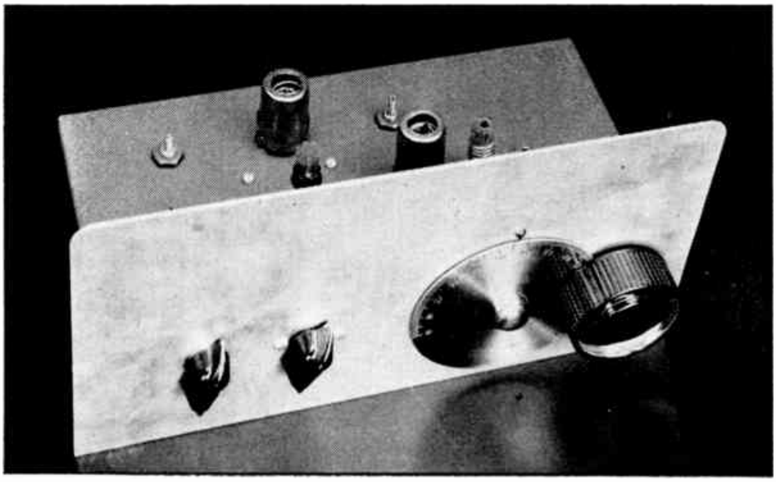

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Fig. 16-19 — A 2-tube converter for 50 and 144 Mc. The vernier dial is a National Type K, with an HRT knob replacing the small one with which the dial is normally equipped. The two other knobs are the i.f. gain control, left, and the mixer tuning condenser.

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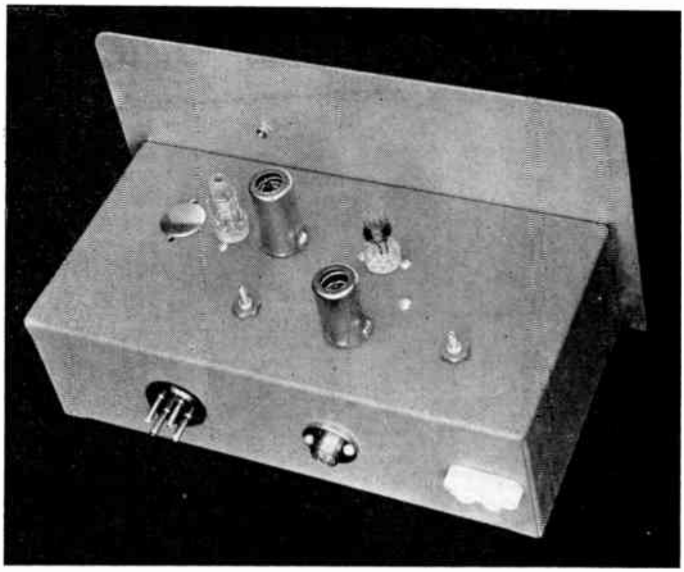

Fig. 16-20 — Rear view of the simple converter. Near the panel, left to right, are the oscillator coil, mixer-oscillator tube, and the mixer grid coil. The 50-Mc. coils are shown. The i.f. amplifier tube is nearer the back of the chassis, with the slug-tuned mixer and i.f. plate coils at either side.

amplifier socket, and the output coil assembly are across the back of the view, from left to right. The power plug, i.f. output fixture, and antenna terminal are on the rear wall in the same order.

Test Procedure

When the assembly and wiring are completed, the oscillator operation should be checked. The power supply should deliver 6.3 volts a.c., at 1 ampere, and 150 volts d.c. at 30 ma., preferably regulated. Insert a milliammeter in series with R_3 and check for oscillation by touching any bare spot in the oscillator plate or grid circuit with a pencil. A change in current indicates oscillation.

The frequency of the oscillator may be checked with an absorption-type wavemeter or Lecher wires. For the 50 Mc. range, the oscillator should tune from 57.4 to 61.4 Mc. in

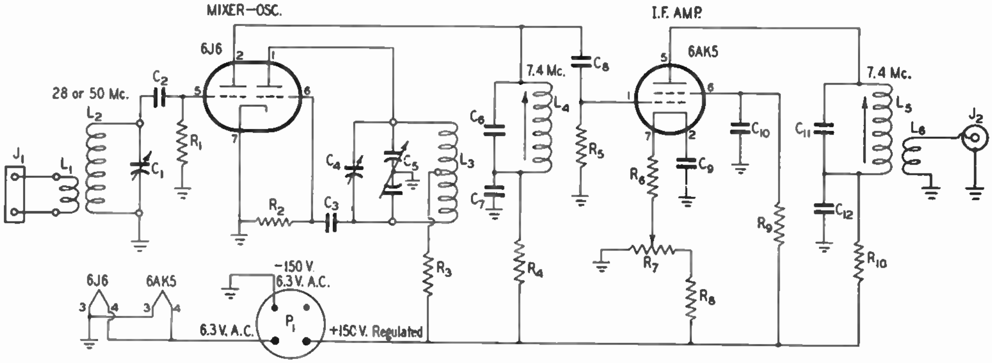

Fig. 16-21 — Schematic diagram of the two-tube converter for 50 and 144 Mc.

C_1 — 15-$\mu\mu$fd. midget variable (Millen 20015).
C_2 — 100-$\mu\mu$fd. mica or ceramic.
C_3, C_8 — 50-$\mu\mu$fd. mica or ceramic.
C_4 — 30-$\mu\mu$fd. air-dielectric padder (Silver 619). Alternative: Ceramic trimmer of similar capacitance, such as Centralab 820-C.
C_5 — Special split-stator variable, 7 plates per section, made from Millen 21935 — see text.
C_6, C_{11} — 68-$\mu\mu$fd. mica or ceramic.
C_7, C_9, C_{10}, C_{12} — 0.01-μfd. disc-type ceramic.
R_1, R_5 — 1 megohm, ½ watt.
R_2 — 10,000 ohms, ½ watt.
R_3, R_4, R_9, R_{10} — 1000 ohms, ½ watt.
R_6 — 220 ohms, ½ watt.
R_7 — 2000-ohm wire-wound potentiometer.
R_8 — 22,000 ohms, 1 watt.
L_1 — 50 Mc: 3 turns No. 22 enamel, close-wound at cold end of L_2.
144 Mc.: 2 turns No. 22 enamel, close-wound at cold end of L_2.
L_2 — 50 Mc: 7 turns No. 22 enamel, ⅜ inch long.
144 Mc.: 2 turns No. 22 enamel, ⅜ inch long.
L_3 — 50 Mc.: 5 turns No. 22 tinned, 7/16 inch long, center tapped.
144 Mc.: ¾ turn No. 12 tinned, center tapped: a 1⅜-inch length of wire formed into a partial circle with an inside diameter of 1/16 inch.
Note: Coils L_1, L_2 and L_3 wound on Millen 69071 forms. For L_4 the form is sawed off and the base only is used.
L_4, L_5 — 23 turns No. 22 enamel, close-wound on National XR-50 slug-tuned forms.
L_6 — 3 turns No. 22 enamel, close wound at cold end of L_5.
J_1 — Antenna terminal (Millen 33102 crystal socket).
J_2 — I.f. output terminal (Jones S-101-D).
P_1 — 4-prong plug (Amphenol 86-CP4).

order to beat with an incoming signal to produce a 7.4-Mc. i.f. (The oscillator is on the *high* side of the signal.) A kick in the oscillator plate current, or a flicker in the voltage-regulator tube in the power supply, can be used to show when the frequency is found with the measuring device.

Set the padder, C_4, so that 57.4 Mc. comes at about 5 divisions in from the maximum-capacity end of the tuning range, and check to see where 61.4 Mc. is found. It should come just inside the minimum-capacity end of the range. If the circuit will not tune to 61.4 Mc. the inductance of L_3 is too low. Move the turns closer together, and reset C_4 as before for 57.4 Mc. If the bandspread is too small, spread the turns and increase the capacitance of C_4 to compensate, for the desired amount of spread, about 90 divisions on the dial.

Next check the 2-meter range. Here the coil must be adjusted in inductance until the oscillator will hit 136.6 Mc. somewhere between the middle and the maximum-capacity end of the tuning range of C_5. The high end, 140.6 Mc. will then appear about 50 to 60 divisions higher on the dial. The oscillator is on the *low* side of the signal on this range. Do not change the setting of C_4 in this process, or it will be necessary to alter the 50-Mc. coil again.

This arrangement (one padder for both bands) does away with the need for padders in the coils themselves, and is worth the added care that must be taken in designing the coils. Somewhat reduced bandspread results on the 144-Mc. band. This can be increased by making the coil smaller, and increasing the value of C_4 accordingly. It will then not be possible to cover the entire 50-Mc. band, but this is no handicap so long as use of the band is concentrated near the low end, as at present.

Once the oscillator is made to tune the proper frequency ranges the converter may be tested in actual reception. Connect the output through a coaxial cable to a receiver tuned to approximately 7.4 Mc. With the converter in operation, there should be an increase in noise as the gain control is turned up. The mixer and i.f. amplifier plate windings can be tuned to the proper frequency merely by adjusting the core screws for maximum noise.

The mixer grid circuit may also be peaked on noise, though care should be taken to see that it is not peaked on the image, 14.8 Mc. away from the signal frequency. If the grid circuit is tuned to the desired frequency there will be a considerable increase in the strength of a signal as the grid condenser, C_1, is tuned through resonance. If the circuit is tuned to the image frequency the noise will peak up, but an amateur-band signal will drop in strength as the noise peak occurs. Tuning the mixer grid circuit shifts the oscillator frequency slightly, so it may be peaked more accurately on noise than when listening to a signal.

A final check of the dial calibration may be made by tuning in signals of known frequency, or by means of an accurate signal generator. Few wavemeters are sufficiently accurate for final calibration by the method outlined earlier. When the desired calibration is attained, the converter is ready for use.

If, in actual operation, trouble is encountered with signals in the 7-Mc. region leaking through, the i.f. can be shifted slightly to tune out the interference. In some instances it may be necessary to put a bottom plate on the chassis. Small changes in intermediate frequency can be made without resetting either the oscillator padder or the i.f. coils. With the i.f. amplifier built into the converter, the setup will have adequate gain for use with almost any receiver. Reception will be nearly as good as with more complex designs, the principal difference being a somewhat higher noise figure (slightly degraded signal-to-noise ratio) in the simpler job. The use of a low-noise r.f. amplifier ahead of the converter (an example is the 6J6 preamplifier of Fig. 16-26) will make possible reception equal to the best obtainable in a converter having a tunable oscillator.

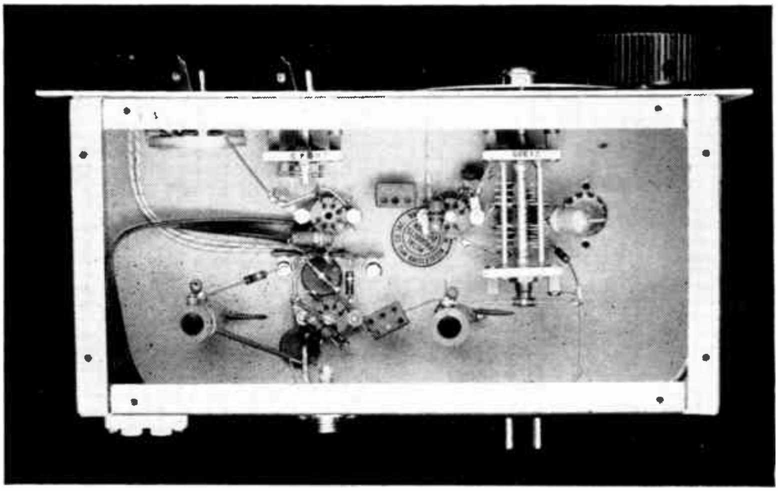

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Fig. 16-22 — Bottom view of the two-band converter. The split-stator condenser at the left is for oscillator tuning. The oscillator coil socket is out of sight above this condenser. At its right is the 6J6 socket. The mixer tuning condenser and grid coil socket are just to the right of the middle of the chassis, with the i.f. coils and tube socket at the rear.

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A Cascode Converter for 220 Mc.

The 220-Mc. converter shown in Figs. 16-23, 16-24 and 16-25 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

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Fig. 16-23 — 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.

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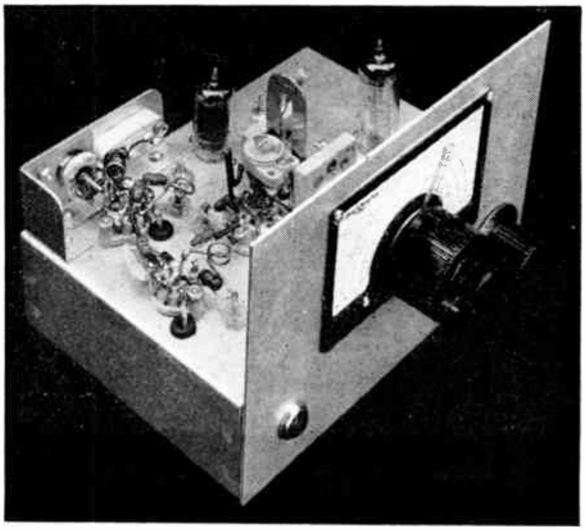

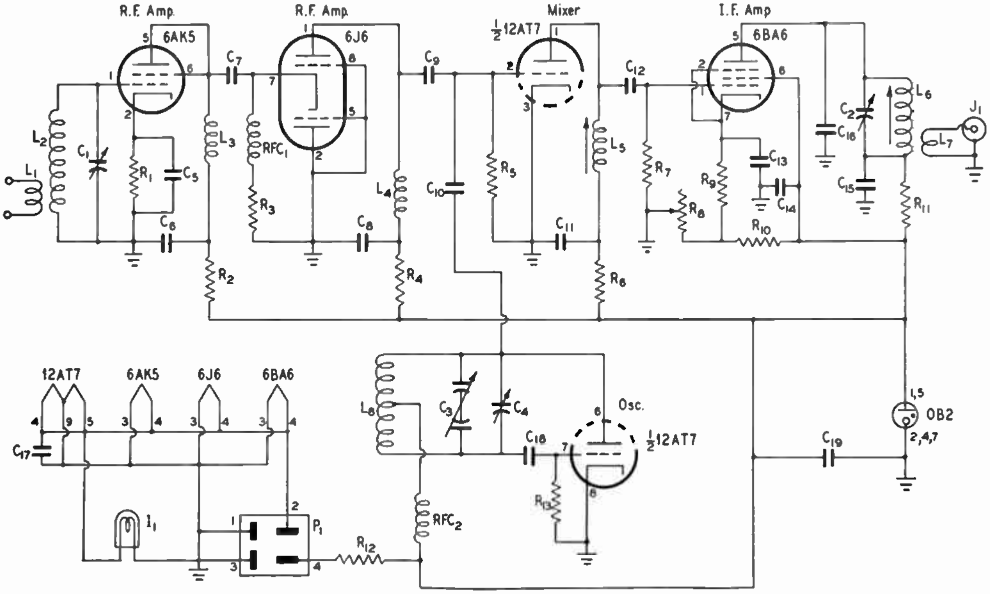

Fig. 16-24 — Schematic diagram of the 220-Mc. converter.

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, ¼-inch diameter, inserted between turns of L_2.
L_2 — 3 turns No. 16 tinned, ¼-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, 1 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.

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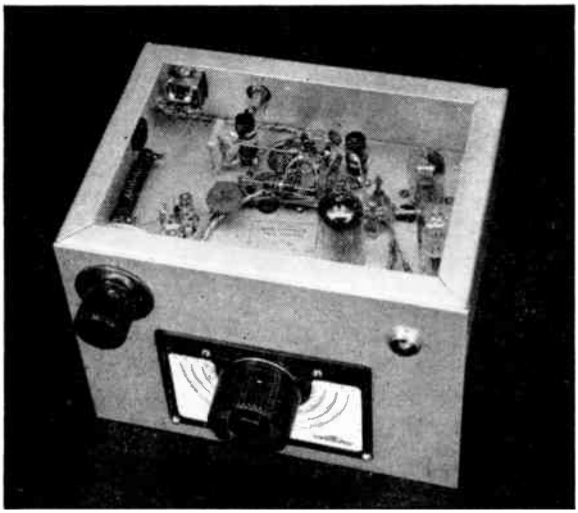

Fig. 16-25 — Under-chassis view of the 220-Mc. converter. The three inverted tubes may be seen at the right. The slug-tuned i.f. coils are at the back of the chassis.

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A 6J6 Preamplifier for 28, 50 and 144 Mc.

The triode preamplifier shown in Figs. 12-26 to 12-29 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.

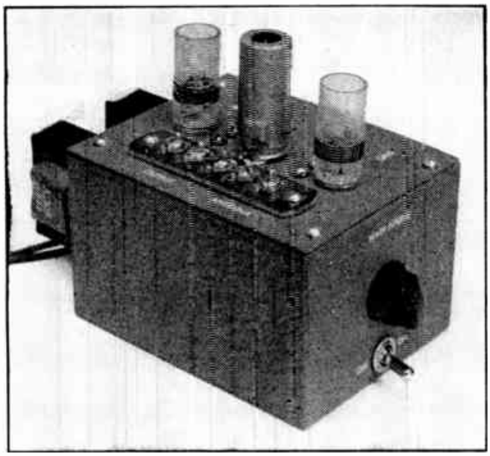

Fig. 16-26 — An r.f. preamplifier for 28, 50 and 144 Mc. The 50-Mc. coils are shown.

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. 12-29 shows the utility box with all power-supply components mounted in place and wired, ready for use.

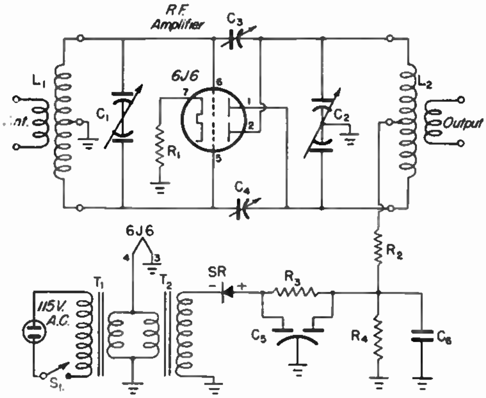

Fig. 16-27 — Schematic diagram of the 3-band r.f. preamplifier.

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-2911).

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

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.

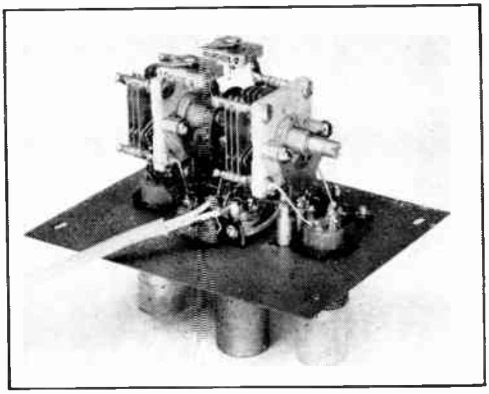

Fig. 16-28 — The r.f. portion of the 3-band preamplifier is mounted on the cover plate of the utility box.

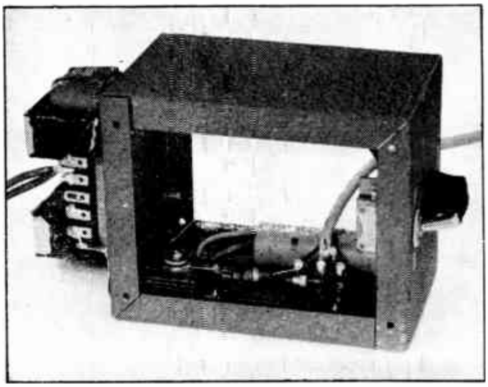

Fig. 16-29 — Power-supply components of the preamplifier are mounted on the walls of the utility box.

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 midpoint 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.

Receivers for 420 Mc.

With crystal control or its equivalent in stability almost standard practice for all bands up through 220 Mc., there is little point in using more bandwidth in the receiver for these frequencies than the minimum needed for satisfactory voice reproduction. Since the sharpest practical receiver will give the best signal-to-noise ratio, all other factors being equal, we will want to keep the bandwidth down in a receiver for 420 Mc. also, but there are other limiting factors at this frequency.

One is the matter of oscillator stability. Even the best tunable oscillators may suffer from vibration and hand-capacity effects at 420 Mc., so it is desirable to use an i.f. bandwidth somewhat

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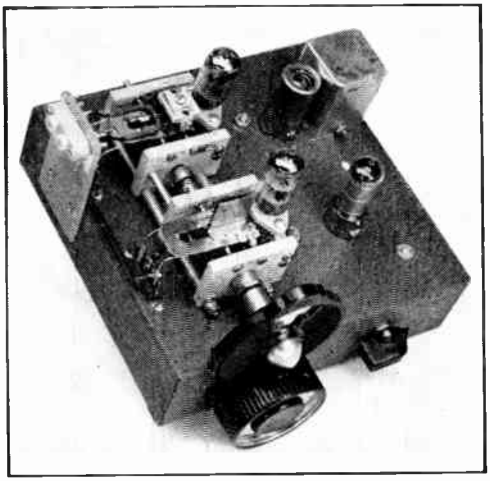

Fig. 16-30 — A converter for 420-Mc. reception. The oscillator section is in back of the vernier dial, with the mixer at the rear. Both use 6J6s in pushpull circuits. The tubes at the right are the 30-Mc. i.f. amplifier, a 6AG5, and a voltage regulator.

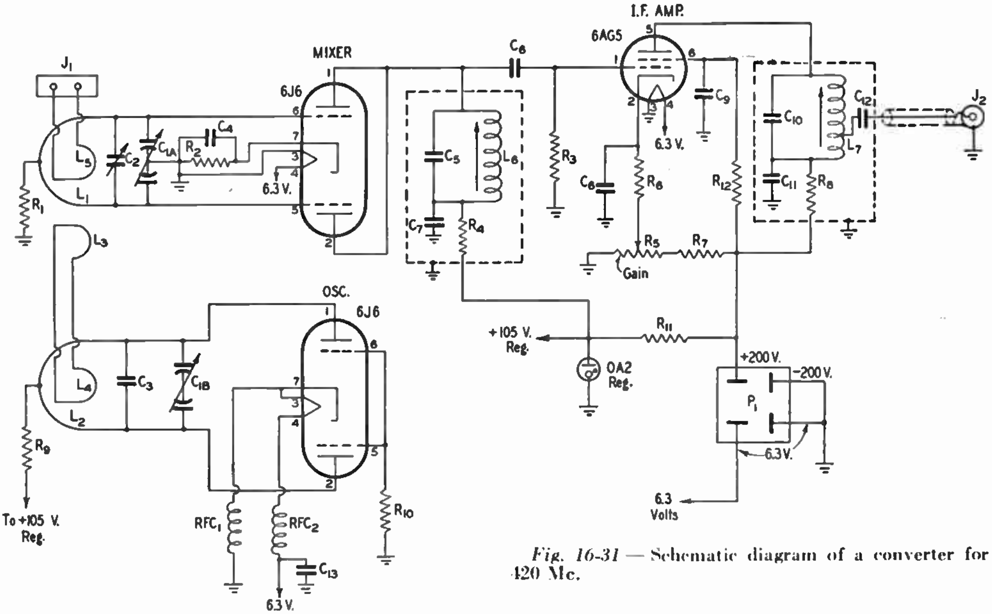

Fig. 16-31 — Schematic diagram of a converter for 420 Mc.

C_1 — Two-section ganged split-stator variable, 6.75-μμfd.-per-section stator to stator (National VHF-2D). One plate may be removed from each section to increase bandspread, if desired.
C_2 — 3–30 μμfd. mica trimmer.
C_3 — Padder capacitance made from two copper plates, ⅞ by 1 inch in size, soldered across terminals of L_2 and C_1. Adjust spacing for band-setting purposes.
C_4, C_7, C_8, C_9, C_{11} — 0.005-μfd. disc ceramic.
C_5, C_{10} — 15-μμfd. ceramic.
C_6 — 50-μμfd. ceramic.
C_{12} — 500-μμfd. ceramic.
C_{13} — 100-μμfd. button by-pass.
R_1 — 470 ohms, ½ watt.
R_2 — 1000 ohms, ½ watt.
R_3 — 1 megohm, ½ watt.
R_4, R_8 — 1000 ohms, ½ watt.
R_5 — 10,000-ohm potentiometer.
R_6 — 68 ohms, ½ watt.
R_7 — 33,000 ohms, 1 watt.
R_9 — 100 ohms, ½ watt.
R_{10} — 3300 ohms, ½ watt.
R_{11} — 2500 ohms, 10 watts.
R_{12} — 33,000 ohms, ½ watt.
L_1, L_2 — U-shaped inductances cut from sheet copper, ⅞ by 1⅞ inches over all. Cut-out portion is ¼ inch wide. Solder directly to flat plates on the tuning-condenser stators, adjusting position of L_2 for proper tracking.
L_3, L_4 — Injection coupling loops of stiff wire, width of L_1 and L_2, and mounted closely under them.
L_5 — Antenna coupling loop of stiff wire 1¾ inches long, coupled closely to L_1.
L_6 — 10 turns No. 24 d.s.c. spaced to fill National XR-50 form.
L_7 — Same as L_6, but tapped at second turn from cold end.
J_1 — Antenna terminal — Millen 33102 crystal socket.
J_2 — Coaxial fitting (Jones S-201).
P_1 — 4-prong power fitting.
RFC_1, RFC_2 — 10 turns No. 22 enameled wire, close-wound on 1-watt resistor.

more than that of the communications type of receiver in 420-Mc. work, even when the signals to be received are of the stable variety. A good compromise in i.f. bandwidth for 420-Mc. reception is that customarily employed in f.m. broadcast receivers, about 200 kc. or so. This is enough so that, with reasonable precautions in oscillator design, a 420-Mc. converter is easily tuned, and even the modulated oscillator type of signal may be received with good quality if the receiver is equipped with f.m. detection, and the transmitter modulation is held to a moderate level.

It is difficult, if not impossible, to build a tunable oscillator for a 420-Mc. converter that will do a completely satisfactory job when communications receiver selectivity is used in the i.f. system. One possible approach is the use of an

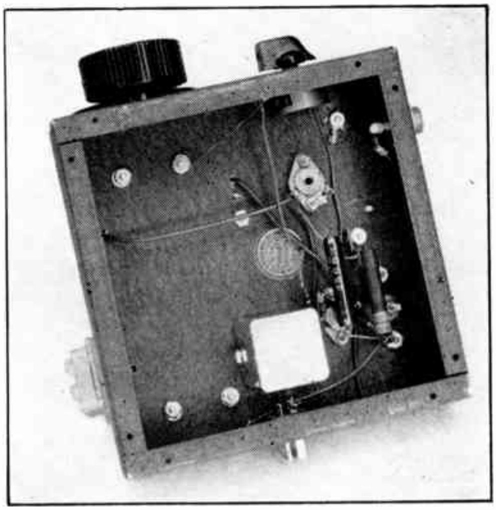

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Fig. 16-32 — Bottom view of the 420-Mc. converter.

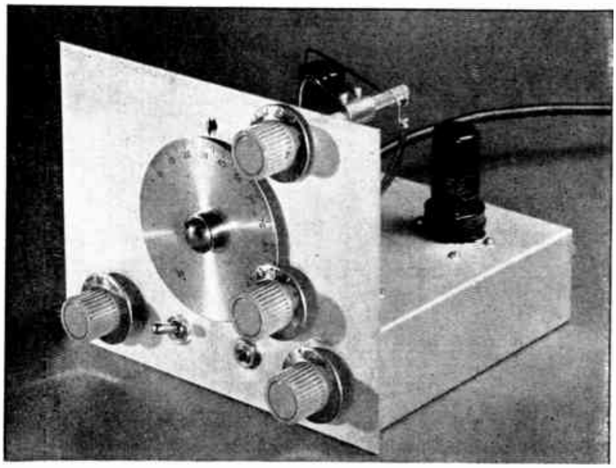

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Fig. 16-33 — A superregenerative receiver for 420 Mc. The two lower controls are for variation of detector voltage (left) and audio gain. The vernier dial is the main tuning and the knob at the top adjusts the antenna coupling.

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oscillator-multiplier system, but this is practical only for tuning a part of the band, unless one or more gang-tuned multiplier stages are used.

The best solution to the stability problem is a crystal-controlled injection source, but this imposes the band coverage problem to a still greater degree. Several crystals and a tunable i.f. system are then required.

Assuming that we have a reasonably stable oscillator and an i.f. bandwidth of 100 to 200 kc., we still may be a long way from satisfactory receiver performance at 420 Mc. Nearly all conventional receiving tubes work poorly, if at all, at this frequency, with the result that the sensitivity and signal-to-noise ratio are much lower at 420 Mc. than would be possible with a comparable tube lineup at 144 Mc.

To date, little success has been had with r.f. amplifier stages using inexpensive receiving tubes, the only suitable tubes for this purpose being the lighthouse and pencil types. These require special tank circuits of the coaxial or flat-plate variety.

Most converters presently used on 420 Mc. thus use only a mixer and an oscillator, followed by one or more i.f. amplifier stages. The i.f. amplifier is a necessity; the output of a 420-Mc. mixer is too low to provide satisfactory performance with the average receiver that might be available for use as an i.f. system.

The mixer may be either a vacuum tube, using more-or-less conventional circuitry, or a crystal diode. Though tube mixers outperform crystal mixers at lower frequencies, in the region around 400 to 600 Mc. there is little choice between the two, and many available tubes may actually be inferior to crystal diodes in 420-Mc. work.

● A 420-MC. CONVERTER

The converter shown in Figs. 16-30 through 16-32 was designed for use in conjunction with communications receivers having provision for wideband f.m. detection. Examples are the S-27, S-36, SX-42, and SX-62. The intermediate frequency is 30 Mc., so it may be used with any receiver covering that range, but best results will be obtained with those having wideband f.m. facilities. It may be used with the SX-43 or with f.m. broadcast receivers, provided that the intermediate frequency is changed to suit the tuning range of the receiver. This would be 42-50 Mc., in the case of receivers for the old f.m. band, or 88-108 Mc. for the present assignment.

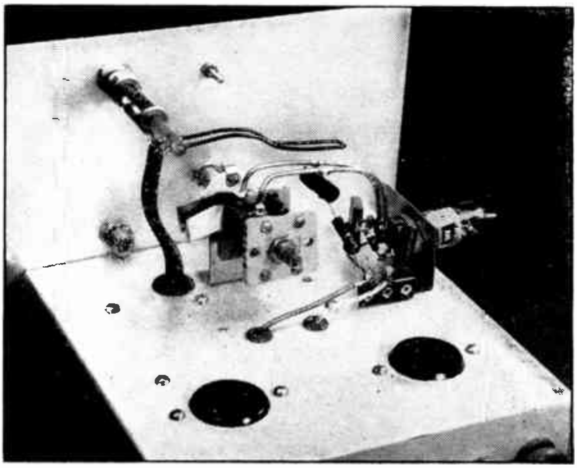

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Fig. 16-34 — Detail view of the 420-Mc. superregenerative receiver. Note the method of varying the antenna coupling. Copper plates attached to the tuning-condenser stators provide a bandset adjustment.

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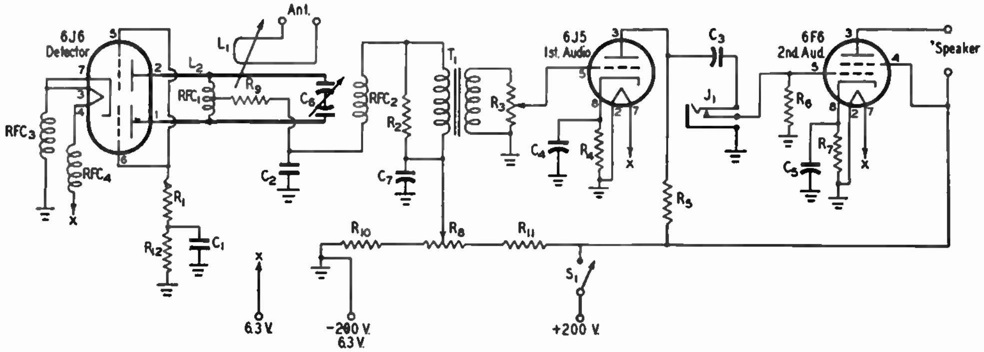

Fig. 16-35 — Schematic diagram of the 420-Mc. superregenerative receiver.

C1 — 470-μμfd. mica.
C2 — 0.0033-μfd. mica.
C3 — 0.01-μfd. tubular.
C4, C5 — 10-μfd. 25-volt electrolytic.
C6 — Miniature split-stator variable, about 4 μμfd. per section. (Millen 21912D), with one rotor plate removed from each section. See text and photo.)
C7 — 0.1-μfd. tubular.
R1 — 3800 ohms, ½ watt.
R2 — 47,000 ohms, ½ watt.
R3 — 0.5-megohm potentiometer.
R4 — 2200 ohms, 1 watt.
R5, R6 — 0.1 megohm, ½ watt.
R7 — 470 ohms, 1 watt.
R8 — 50,000-ohm potentiometer.
R9 — 2200 ohms, 1 watt.
R10, R11 — 47,000 ohms, 1 watt.
R12 — 1.5 megohms, ½ watt.
L1 — Hairpin loop No. 14 enameled wire, same spacing as *L2*. Connect to antenna terminals by means of 300-ohm line.
L2 — 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 *L1* and *L2*.)
J1 — Closed-circuit jack.
RFC1 — 19 turns No. 20 enameled wire, 3/16-inch inside diameter, ⅞ inch long, center-tapped.
RFC2 — 10-mh. r.f. choke.
RFC3, RFC4 — 12 turns No. 20 enameled wire, 3/16-inch inside diameter, ¾ inch long.
S1 — S.p.s.t. toggle switch.
T1 — Interstage audio transformer.

Such a converter may be used for reception of amateur television signals in the 420-Mc. band by adjusting the intermediate frequency to a television channel that is not used locally. A channel in the low band is recommended.

The mixer and oscillator stages use 6J6s, with gang-tuned pushpull circuits. A 30-Mc. i.f. amplifier is included, as the gain of most receivers at 30 Mc. is insufficient for best reception. The i.f. stage uses a 6AG5, which works well at this frequency, but if the i.f. is to be shifted to the 90-Mc. region it would be well to use the cascode circuit in the i.f. amplifier, for adequate gain and low-noise characteristics. Details of the cascode amplifier will be found earlier in this chapter. Plate voltage for the oscillator and mixer is maintained at 105 volts by means of an OA2 regulator tube.

The tuning condenser is a ganged unit especially designed for v.h.f. service. The mixer and oscillator inductances, *L*1 and *L*2, are cut from sheet copper in U shape, and soldered directly to the stator assemblies in the tuning condenser. The 6J6 tube sockets are mounted on brackets supplied with the condenser assembly, permitting connections to be made without leads other than the socket lugs themselves. Padder capacitance for the oscillator is supplied by two copper plates, also soldered directly to the stator terminals.

● A SUPERREGENERATIVE RECEIVER FOR 420 MC.

Where simplicity, low cost, and low power consumption are important a superregenerative receiver may be useful. Such a receiver is shown in Figs. 16–33 through 16–36. A 6J6 is used as a pushpull detector with a halfwave line in the plate circuit, followed by two metal-tube audio stages. Tuning is done by means of a small split-stator condenser across the open end of the line.

Adjustment of the antenna coupling for best reception may be quite critical, so a convenient

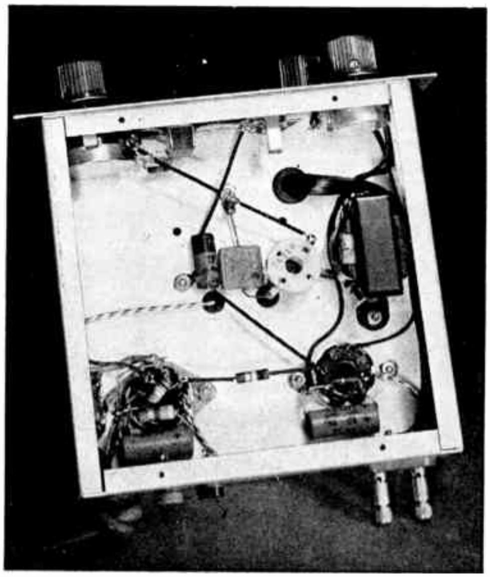

Fig. 16-36 — Bottom view of the 420-Mc. receiver. Loudspeaker terminals are at the lower left. At the right are the antenna terminals, from which a length of 300-ohm line runs up through the chassis to the antenna coupling loop.

means for varying the position of the coupling loop is provided. Regeneration is controlled by varying the detector plate voltage. In the rear-view photograph, Fig. 16–34, the audio tubes have been removed to show the details of the halfwave line and the antenna coupling assembly. It will be seen that the line is mounted horizontally, with each end bent over to reduce the space required. Small plates of sheet copper are soldered to the stator plate of the tuning condenser, and these are used to form a band-setting padder capacitance.

The frequency of the superregenerative detector may be checked with Lecher wires, or by means of a calibrated wavemeter. When it has been adjusted to tune approximately the correct frequency the optimum point for connection of the center-tapped r.f. choke, RFC_1, should be found. With a temporary connection near the midpoint of the line, touch a pencil lead along the line until the least effect on the operation of the detector is noted. This is the point at which the B-plus should be fed into the line.

All adjustments on a superregenerative detector interlock somewhat, so it is advisable to go over them several times for best results. The most sensitive operating point will be with the tightest coupling to the antenna that will permit superregeneration. If the antenna does not pull the detector out of oscillation at any position of L_1, back of the detector plate voltage until just before superregeneration ceases.

CHAPTER 17

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 in v.h.f. work. Crystal-controlled transmitters and receivers having the minimum bandwidth necessary for voice communication make it possible for hundreds of stations to operate without undue interference in a band that formerly appeared crowded when occupied by a dozen or less stations using broadband receivers and unstable transmitters.

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., but best performance is obtained with the "lighthouse", "pencil tube," or coaxial-electrode types built especially for u.h.f. applications, and requiring specially-designed tank circuits.

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 narrowband FM can provide effective v.h.f. communication, their use being particularly advantageous in congested areas where the freedom from interference to broadcast reception they enjoy may permit operation when an amplitude-modulated transmitter of any power would be a constant source of trouble.

Transmitter Technique

The low-power stages of a transmitter for the v.h.f. bands need not be greatly different in design from those used for lower bands, and many of the ideas in chapter 6 may be used to good advantage in the initial stages of the v.h.f. rig. The constructor has the choice of starting at some lower frequency, usually around 6, 8 or 12 Mc., multiplying to the operating frequency in one or more additional stages, or he can use a higher initial frequency and thus reduce the number of multiplier stages required or eliminate them entirely. The first approach has the virtue of employing low-cost crystals, and it usually results in better stability when methods other than crystal control are used, but high-frequency crystals may effect a considerable economy in power consumption, an important factor in portable or emergency-powered gear.

CRYSTAL OSCILLATORS

Crystal oscillator stages for v.h.f. transmitters may make use of any of the circuits shown in Chapter 6, when crystals up to 12 Mc. are employed, but certain variations are helpful for higher frequencies. Crystals for 12 Mc. or higher are usually of the overtone variety. Their frequency of oscillation is an approximate multiple of some lower frequency, for which the crystal is actually ground. Thus 24-Mc. crystals commonly used in 144-Mc. work are 8-Mc. cuts, specially treated for overtone characteristics. Until recent years such crystals were tricky in operation and subject to excessive drift if operated at high crystal current. The overtone crystals now being supplied are approximately as stable as those designed for fundamental operation, and they are easy to handle in properly-designed circuits.

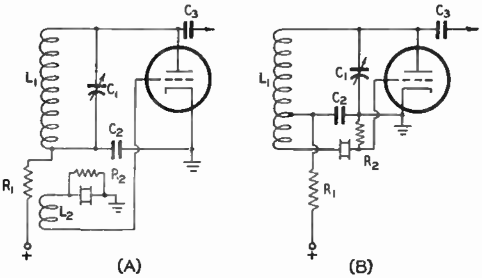

Fig. 17-1 — Regenerative crystal oscillator circuits for v.h.f. use. Feedback is controlled by the position of L_2 with respect to L_1 in A, or by the position of the tap on L_1 in B. Constants below are for 24 to 27 Mc.

C_1 — 50-μμfd. variable

C_2 — 0.005-μfd. ceramic or mica.

C_3 — 25-μμfd. ceramic or mica.

R_1 — Decoupling resistor, 1000 to 5000 ohms, carbon.

R_2 — Grid leak, to suit tube used.

L_1 (A) — 18 turns No. 18, ½-inch dia., 1¼ inches long.

L_2 (A) — 3 turns similar to A, mounted on same axis, about ⅛ inch apart.

L_1 (B) — 14 turns No. 18, ½-inch dia., 1 inch long. Tap at about 4½ turns (see text).

Best results are usually obtained with overtone crystals if some regeneration is added. This makes for easy starting under load and greater output than would be obtainable in a simple triode or tetrode circuit. Two regenerative circuits, with constants for 24- or 25-Mc. crystals, are shown in Fig. 17-1. Triodes are shown, but the same arrangement may be used with tetrode or pentode tubes. The important point in either case is the amount of regeneration, controlled by the position of the feedback winding, L_2, in Fig. 17-1-A, or the position of the tap on L_1 in B. There should be only enough feedback to assure easy crystal starting and satisfactory operation under load; too much will result in random oscillation not under the control of the crystal.

Overtone operation is possible with standard fundamental-type crystals, using the circuits of Fig. 17-1. Practically all will oscillate on their third overtones, and fifth and higher odd overtones may be possible. Adjustment of regeneration is more critical, however, as the crystals are not ground for overtone characteristics. It should also be noted that the frequency may not be an exact multiple of that marked on the crystal holder, so care should be used in working with crystals that are near to a band edge.

Crystals are now available for frequencies up to around 100 Mc. They are somewhat more expensive than those for 25 Mc. and lower, however, so they have not been used widely in amateur work, except where a saving in power is important. Use of 50-Mc. crystals is made occasionally as a means of preventing radiation of the harmonics of lower frequency crystals that might cause interference to television reception.

FREQUENCY MULTIPLIERS

Frequency multiplying stages in a v.h.f. transmitter follow standard practice, the principal precaution being arrangement of components for short lead length and minimum stray capacitance. This is particularly important at 144 Mc. and higher. To reduce the possibility of radiation of oscillator harmonics on frequencies that might interfere with television or other services, the lowest satisfactory power level should be used. Low powered stages are easier to shield or filter, in case such steps become necessary.

Common practice in v.h.f. exciter design is to make the tuned circuits capable of operation over the whole range from 48 to 54 Mc., so that the output stage can drive either a 50-Mc. amplifier or a tripler from 48 to 144 Mc. Tripling is often done with pushpull stages, particularly when the output frequency is to be 144 Mc. or higher. The output capacitances of the tubes in such a circuit are in series, permitting a better L/C ratio than is possible with single-ended circuits.

AMPLIFIERS

Most transmitting tubes now used by amateurs will work on 50 Mc., but for 144 Mc. and higher the tube types are limited to those having low input and output capacitances and compact physical structure. Leads must be as short as possible, and soldered connections should be avoided in high-powered circuits, where heating

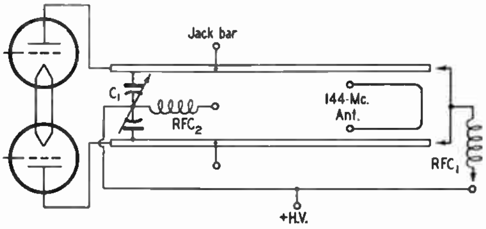

Fig. 17-2 — An efficient two-band tank circuit for 50 and 144 Mc. For operation on 144 Mc. the shorting bar is plugged into the end of the line. For 50 Mc. a suitable tank coil is plugged into the jack bar. The line then serves merely as a pair of plate leads. RFC_1 is a 144-Mc. choke; RFC_2 a 50-Mc. choke. The split-stator variable, C_1, tunes either circuit.

may be great enough to reach the melting point of the solder used.

Plug-in coils and their associated sockets or jack bars are generally unsatisfactory for use at 144 Mc. and higher because of the stray inductance and capacitance they introduce. One way around this trouble is the dual tank circuit shown in Fig. 17-2. Here the tank circuit for 144 Mc. is a conventional tuned line, with its shorting bar made removable by plugs or clips. When the stage is to be used on another band the shorting bar is removed and a coil is plugged into the jack bar, the line then serving as a pair of plate leads. Such an arrangement will operate as efficiently on 144 Mc. as if it were designed for that band alone, yet it can be made to work properly on any lower band.

At 220 Mc. and higher it may be necessary to employ halfwave lines as tuned circuits, as shown in Fig. 17-3. Here the tuning capacitance, instead of being connected directly in parallel with the output capacitance of the tube, is at the far end of a halfwave line. Plate voltage is fed into the line near the middle, at the point where the r.f. voltage is lowest. The proper point can be located by first operating the stage with the voltage fed in near the middle of the line, and then touching a pencil point along the line to locate the spot where the least effect on the grid or plate current is noted. This check should be made with the pencil in an insulating mount, if dangerous values of plate voltage are used.

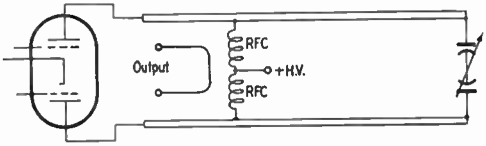

Fig. 17-3 — Halfwave line tank circuit, for use at 220 or 420 Mc., where tube and circuit capacitances prohibit the use of an ordinary tuned circuit. Plate voltage is fed into the line at the point of lowest r.f. voltage (see text).

Neutralization of triode amplifiers for 50 and 144 Mc. can follow standard practice, but the stray inductance and capacitance introduced by the neutralizing circuits may be excessive for 220 Mc. and higher. In such instances grounded-grid amplifiers may be used as shown in Fig. 17-4. Driving power is applied to the cathode circuit, with the grid acting as a shield. Grounded-grid amplifiers are stable, but they require high driving power. Some of the drive appears in the output, so both the driver and amplifier must be modulated when amplitude modulation is used. For this reason the grounded-grid amplifier is used mainly for f.m. applications.

Tetrode and pentode amplifiers may operate without neutralization, but it is advisable to make plans for neutralization in the original layout, as it is often needed. With such tubes as the 829 or 832 enough neutralizing capacitance can be obtained by running short lengths of stiff wire up through the chassis alongside the tube plates, crossing them over to the opposite grid terminals below the chassis. Neutralization is adjusted by trimming or bending the wires.

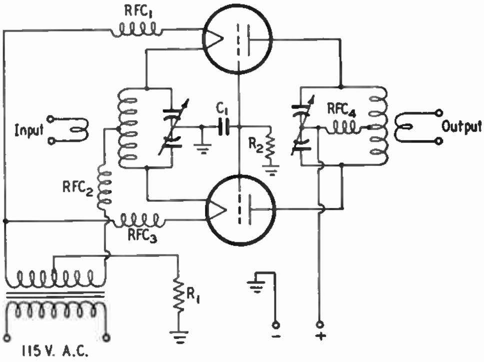

Fig. 17-4 — Grounded-grid r.f. amplifier. Driving voltage is fed into the cathode circuit, with the control grids maintained at ground potential.

Instability may show up in tetrode amplifiers as the result of ineffective screen by-passing, in which case conventional cross-over neutralization will accomplish little or nothing. The solution lies in series-resonating the screen circuits to ground, as shown in Fig. 17-5. A small split-stator variable can be used for C_1 and C_2 if the layout is completely symmetrical. The r.f. choke and condenser values vary with frequency, so screen neutralization is essentially a one-band device.

FREQUENCY MODULATION

Though f.m. has not enjoyed great popularity in v.h.f. operation, probably because of lack of suitable receivers in most v.h.f. stations, its possibilities should not be overlooked, particularly for the higher bands. At 420 Mc., for instance, the efficiency of most amplifiers is so low that it is often difficult to develop sufficient grid drive for proper a.m. service. With f.m. any amount of grid drive may be used without affecting the audio quality of the signal, and the modulation process adds nothing to the plate dissipation. Thus considerably higher power can be run with f.m. than with a.m. before damage to the tubes develops or the signal is of poor quality.

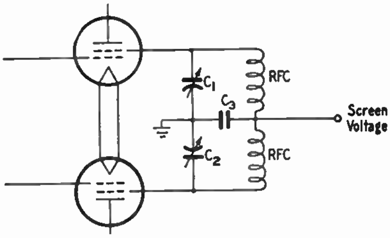

Fig. 17-5 — Tuned screen circuit for stabilizing a v.h.f tetrode pushpull amplifier. C_1 and C_2 may be the two halves of a split-stator variable condenser, if the circuit is symmetrical electrically. The r.f. choke and condenser values vary with frequency, making this form of neutralization essentially a one-band device. C_3 should be about 0.001 μfd. for v.h.f. applications.

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. 17-6 to 17-12 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. 17-8. The 6AR5 Tri-tet oscillator employs a fixed-tuned cathode circuit, C_8L_3. The plate circuit, C_1L_4, 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. 17-8. 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

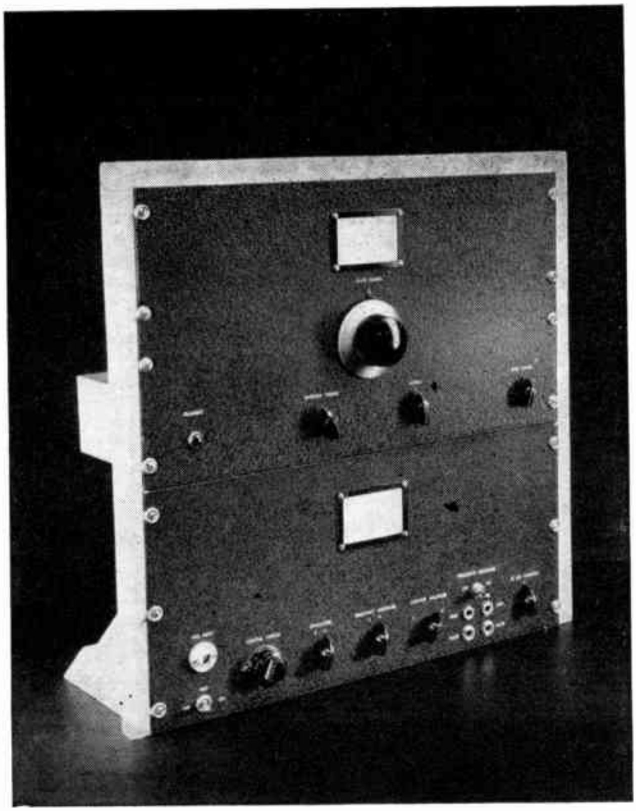

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Fig. 17-6 — A complete 400-watt transmitter for 50 and 144 Mc.

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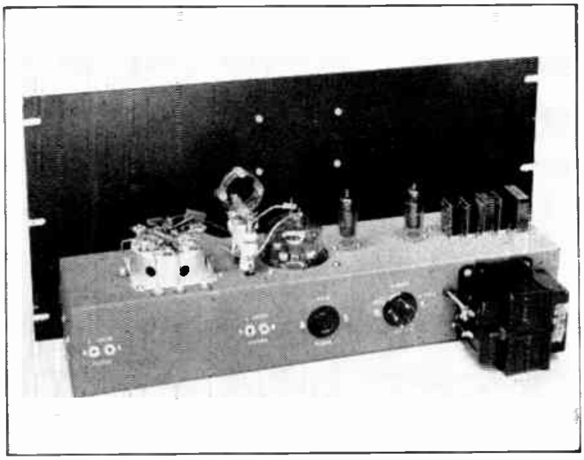

Fig. 17-7 — A rear view of the 50- and 144-Mc. exciter. Across the top of the chassis, from right to left, are the crystal sockets, the oscillator and doubler tubes, the 832A amplifier-tripler and its plate coil, and the inverted 144-Mc. amplifier stage. Crystal sockets, used as r.f. output terminals, 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_4, 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. 17-13 to 17-15, 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 17-I will assist in the selection of a crystal for

TABLE 17-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	114
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.

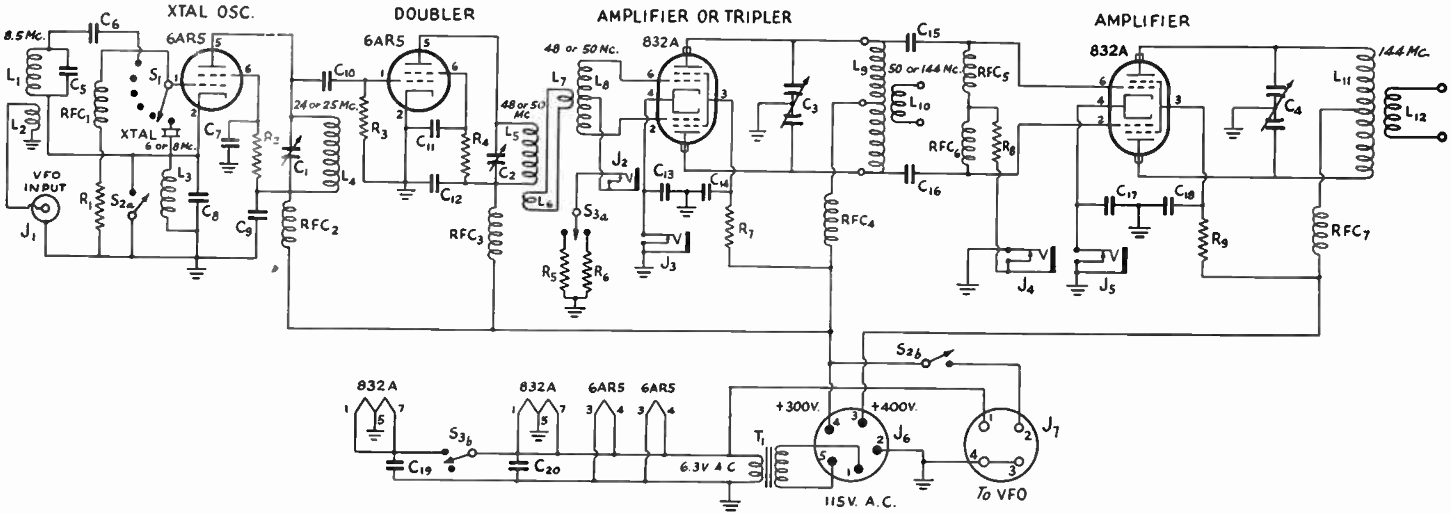

Fig. 17-8 — Circuit diagram of the 50-144 Mc. exciter.

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., ⅜ 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, ⅞ inch long, ⅝-inch diam.
L_4 — 10 turns No. 20 tinned, ⅝ inch long, ⅝-inch diam.
L_5 — 5 turns No. 20 tinned, $\frac{5}{16}$ inch long, ⅝-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., ⅝ 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, ⅞ 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, ⅝-inch diam., wound in two sections with two turns each side of center tap and a ⅜-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 jack
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.

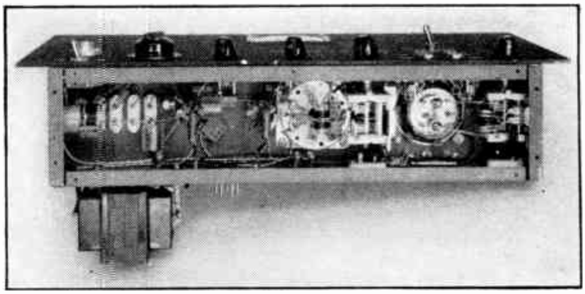

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Fig. 17-9 — Bottom view of the v.h.f. exciter. The VFO input coil is at the left end of the chassis. Plate coils for the oscillator, the doubler and the 144-Mc. amplifier circuits are mounted on the tuning condensers. The grid coil for the amplifier-tripler stage is mounted on the tube-socket terminals.

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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.

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Fig. 17-10 — 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.

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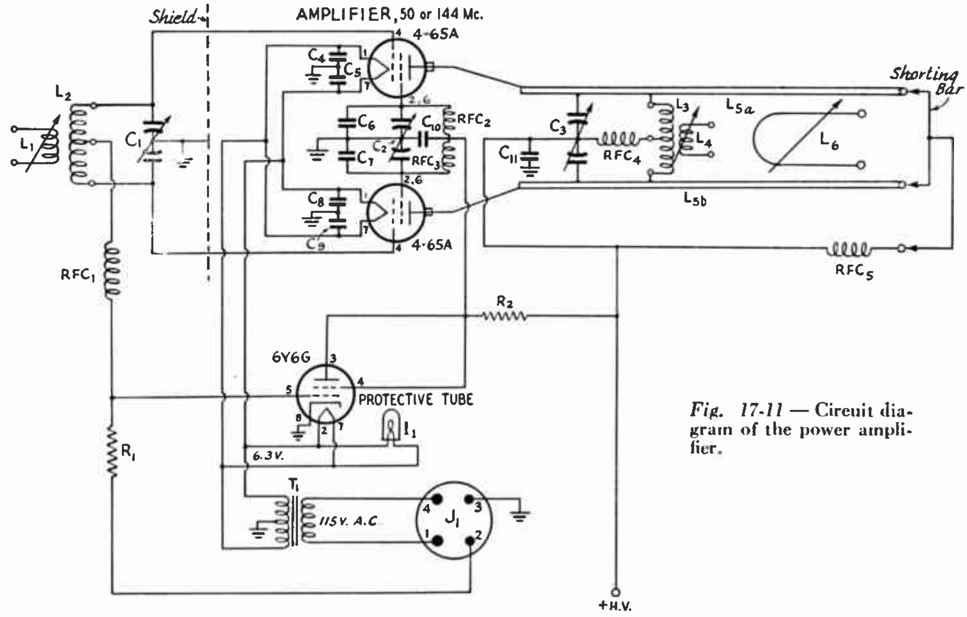

Fig. 17-11 — Circuit diagram of the power amplifier.

C_1 — 6-μμfd.-per-section variable (Millen 21906D).
C_2 — 50-μμfd.-per-section (Bud LC 1662).
C_3 — 35-μμfd.-per-section with high-voltage coupling; see text for information on removing plates. (National TMH-35D.)
C_4, C_5, C_8, C_9 — 470-μμfd. midget mica.
C_6, C_7 — 0.0022-μfd. mica.
C_{10} — 0.001-μfd. mica.
C_{11} — 500-μμfd. 5000-volt mica.
R_1 — 5000 ohms, 10 watts.
R_2 — 30,000-ohm 200-watt adjustable: two 100-watt resistors in series.
L_1 — 50 Mc.: 5 turns No. 21 tinned, ½-inch diam., turns spaced wire diam.
— 144 Mc.: 1 turn No. 14 wire, hairpin shape, 1⅛ inches long, ⅝-inch diam. at open end.
L_2 — 50 Mc.: 6 turns each section, No. 20 tinned, ½-inch diam. (B & W Miniductor No. 3007). Space sections 5⁄16 inch apart.
— 144 Mc.: Same as 144-Mc. L_1.

NOTE: L_1 and L_2 mounted on National type PB-16 plugs.

L_3 — 4 turns of ⅛-inch o.d. copper tubing, 1⅝-inch diam., wound in two sections with two turns each side of center tap and a ¾-inch space at center, turns spaced ⅛ inch.
L_4 — 3 turns No. 12 enam., 1⅝-inch diam., turns spaced wire diam.
L_{5A}, L_{5B} — ½-inch o.d. copper tubing, 10½ inches long, spaced 1⅛ inches on centers.
L_6 — 1 turn of ⅛-inch copper tubing, hairpin shape, 3 inches long, 1¼-inch diam. at open end.
I_1 — 6.3-volt pilot-lamp assembly.
J_1 — 4-prong male receptacle.
RFC_1 — 1-μh. r.f. choke.
RFC_2, RFC_3, RFC_4 — 7-μh. r.f. choke (Ohmite Z-50).
RFC_5 — 1.8-μh. r.f. choke (Ohmite Z-144).
T_1 — Filament transformer: 6.3 volts, 8 amp.

coil assembly into the jack bar. Individual antenna coupling adjustments are used, the one for 144 Mc. being adjustable from the front panel.

The circuit diagram of the push-pull amplifier is given in Fig. 17-11. Excitation for the amplifier is link coupled to a conventional split-stator grid circuit. A 6Y6G protective tube holds the plate dissipation to a safe level when the excitation is removed. The tubes require no neutralization at 50 Mc. At this frequency the screen grids are by-passed by condensers C_6 and C_7. Shielding to prevent external coupling between the grid and the plate circuit is provided for by an aluminum partition.

On 144 Mc. it is necessary to series-tune the screens to ground by means of C_2, placing the screens effectively at ground potential. This is a frequency-sensitive adjustment, however, and the stability of the amplifier should be checked after making large changes in operating frequency.

Construction

The 3 × 7 × 17-inch chassis and the 10½ × 19-inch panel are held together by the pilot-lamp assembly and three shaft bearings. The latter are for the 144-Mc. output link and the tuning condensers for the screen and grid circuits. The lamp jewel and the three control knobs may be seen from left to right in the front view of the complete transmitter.

The rear view of the amplifier, Fig. 17-10, shows the grid coil mounted on a National type XB-16 socket to the left of the shield partition. An XB-15 socket is mounted on 3-inch stand-off insulators between the 4-65A tubes and the plate tuning condenser. The condenser is mounted on 2½-inch insulators in an inverted position. The minimum capacitance of the plate condenser was reduced by removing two stator and two rotor plates from each section. A feed-through insulator for the high-voltage d.c. lead is mounted directly below the plate-coil socket.

The 144-Mc. lines are supported by the tuning condenser and a piece of ¼-inch polystyrene. Plate clips for 9/16-inch caps are reduced in diameter and used for contact with the rods at the tube and condenser positions. The condenser should have the clips bolted to the left-hand stator terminals as seen from the rear view. This will allow the condenser shaft to be centered on the panel and the connection to the lines will be at a point four inches in from the plate ends. Shield braid, ½ inch wide, is used between the clips at the open ends of the lines and the heat-radiating caps of the tubes.

Aluminum plates equipped with panel bearings for the control shaft of the 144-Mc. output link are mounted on the front and the rear frames of the plate tuning condenser. The swinging link is made by twisting the open ends of the loop around a 5-inch length of ¼-inch polystyrene rod. The turns around the rod should be shorted out by soldering, and since this operation softens the rod, the link and rod will be firmly joined together. A piece of ⅛-inch polystyrene cemented across the closed end of the loop prevents accidental contact with the plate lines. A Millen type 38602 Quartz Q washer at the rear of the shaft and a homemade pulley at the front prevent the control shaft from slipping out of the bearings.

The grid tuning condenser is mounted on an aluminum bracket and the screen condenser is supported by metal posts as shown in the bottom view. Copper strip is used for joining the two screen prongs of each tube socket and for connection to the two variable condensers. Each tube has the mica by-pass condensers and a section of the variable screen condenser returned to a common point on the socket. The 0.001-μfd. by-pass for the screen-circuit r.f. chokes is returned to ground in between the two sockets. The pulley and the dial cord for the swinging link are at the front of the chassis.

Testing the Complete Rig

Although the amplifier described may be operated at full ratings (540 watts) for c.w. work, it is recommended that the input be kept to 400 watts or less when plate modulation is used. This value includes the power taken by the screen grids. For all-round operation a power-supply output of 1000 to 1500 volts at approximately 350 ma. is recommended. Higher voltages may be used but forced-air cooling of the tubes may be required. The amplifier may also be operated efficiently at supply voltages as low as 600, provided that the screen is maintained at approximately 250 volts.

At 50 Mc. the amplifier can be tested in the same way as any low-frequency amplifier. The usual test for self-oscillation may be made with the filament and plate voltage applied, with the excitation removed and with the protective tube in place. Proper operation is indicated by the absence of grid-current as the grid and the plate tuning condensers are rotated. The protective tube should limit the d.c. input to no more than the maximum plate-dissipation rating. The limiting effect will be determined by the supply voltage, but total input should be well below 150 ma. at 1500 volts or less.

With the unit described earlier furnishing excitation, the grid current for the amplifier should be approximately 35 ma. before the high voltage is turned on. A 300-watt lamp coupled to L_4 may be used as a dummy load for the power-output test. A cathode current of 320 ma. is correct for operation with a 1000-volt supply, and 310 ma. is correct with 1500 volts on the plates. The grid current should be at least 25 ma. and the screen potential should be about 250 volts when the amplifier is fully loaded.

The amplifier is tested for 144-Mc. operation with the 50-Mc. plate coil removed and with the shorting bar across the resonant lines. The one addition to the test procedure outlined above is adjustment of the screen tuning condenser, C_2. After applying filament voltage and excitation, the condenser is adjusted for minimum feed-through as indicated by a sensitive rectifier-type wavemeter coupled to the plate lines. A second method is to remove the excitation, apply the plate voltage and then tune for zero grid current. The setting of the screen control is very critical, but with care a position can be found which will hold over most of the 144-Mc. band. The most accurate way of setting the adjustment is to try for a position where maximum grid current and minimum plate current occur at the same setting of the plate tuning condenser.

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Fig. 17-12 — A bottom view of the power amplifier. Tuning condensers for the grid and screen circuits are to the left and right of the shielded tube sockets. The filament transformer is at the upper right-hand end of the chassis. The two large resistors drop the voltage to the 4-65A screens.

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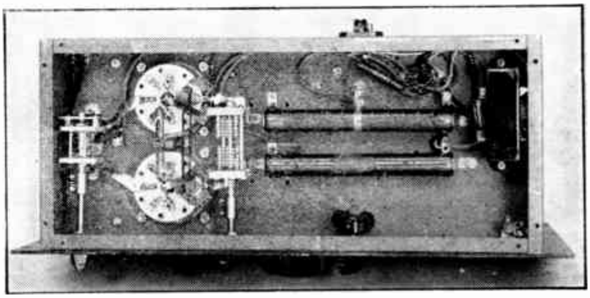

A V.H.F. Man's VFO

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. 17-13–17-15, 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

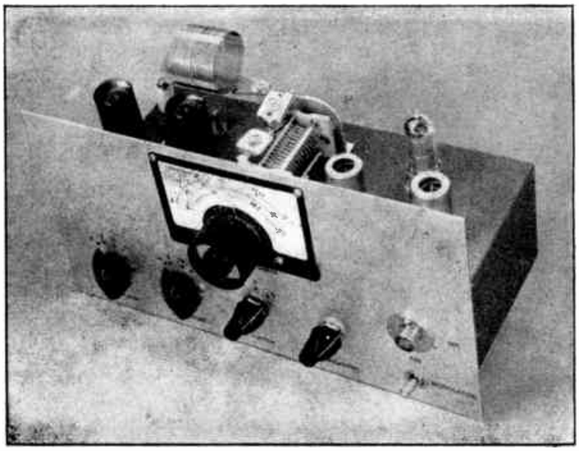

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Fig. 11-13 — Panel view of the v.h.f. VFO with NFM modulator.

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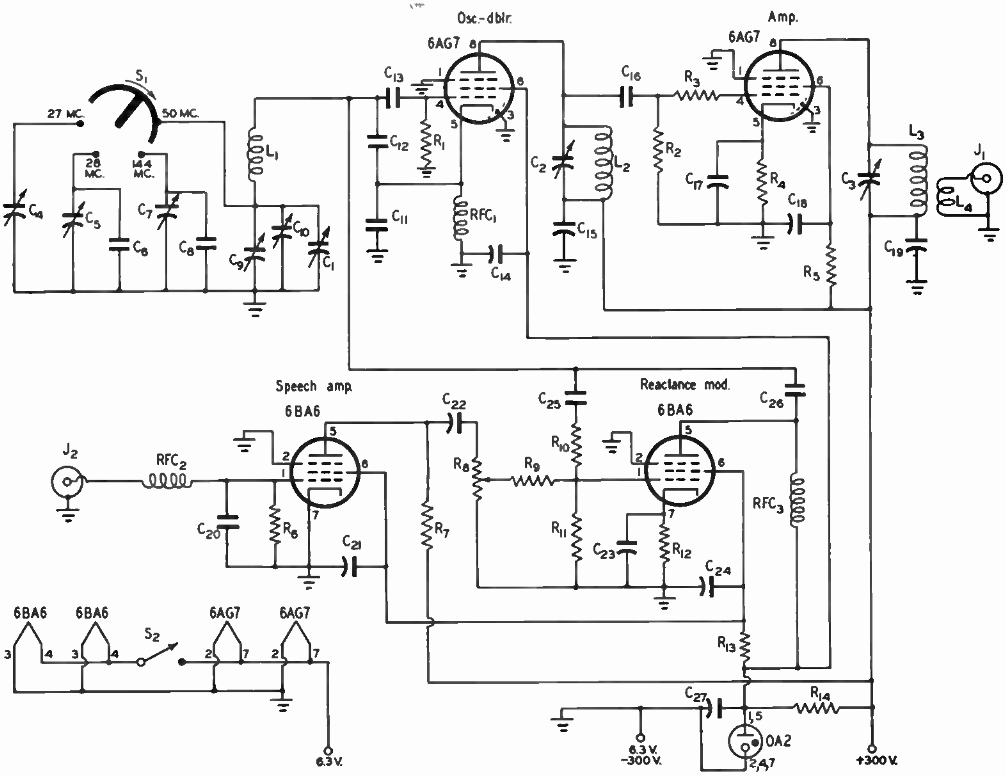

Fig. 17-14 — Circuit diagram of the NFM control unit for v.h.f. use.

C_1 — 35-μμfd. variable, double spaced (Millen 21935).
C_2, C_3 — 100-μμfd. variable (Millen 20100).
C_4, C_5, C_7, C_9, C_{10} — 2–30-μμfd. ceramic trimmer (Millen 27030).
C_6 — 33-μμfd. silver mica.
C_8 — 10-μμfd. silver mica.
C_{11}, C_{12} — 680-μμfd. silver mica.
C_{13} — 68-μμfd. 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-μμfd. mica.
C_{25}, C_{26} — 47-μμfd. 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. 17-13, 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. 17-15. 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. 17-14, 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.

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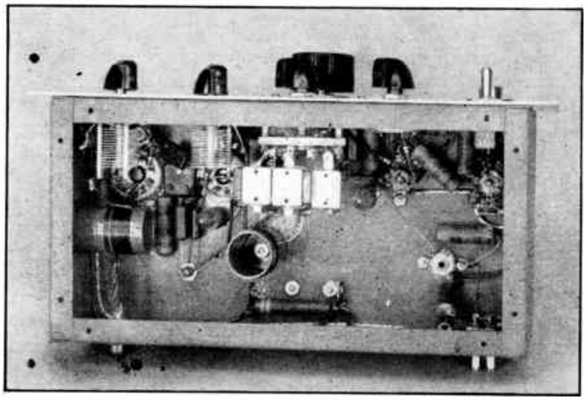

Fig. 17-15 — **Bottom view of the VFO.**

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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. 17-16, 17-17 and 17-18 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 an amplifier such as that in Fig. 17-13.

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

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Fig. 17-16 — 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.

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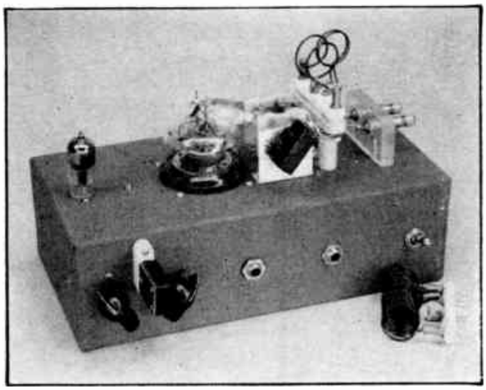

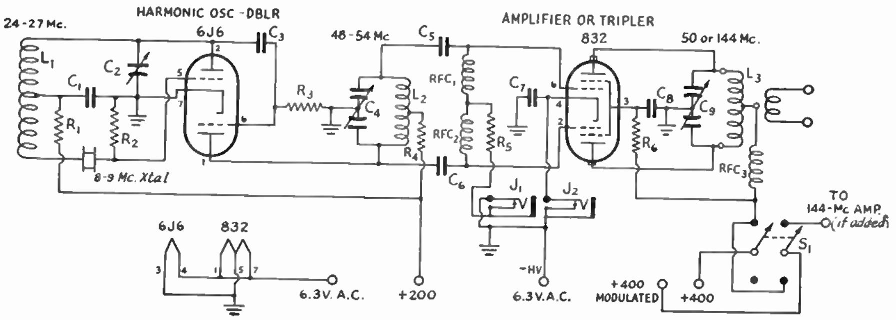

Fig. 17-17 — Schematic diagram of the 2-tube v.h.f. rig. The power-switching arrangement shown provides for later addition of a 144-Mc. amplifier stage.

C_1 — 680-μμfd. mica.
C_2 — 50-μμfd. variable.
C_3 — 15-μμfd. ceramic.
C_4 — 20-μμfd.-per-section split-stator, made by sawing the stator bars of a Millen 21050 and removing center stator and back rotor plates.
C_5, C_6 — 75-μμfd. ceramic.
C_7, C_8 — 500-μμfd. ceramic.
C_9 — 6-μμfd.-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 1½ 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.

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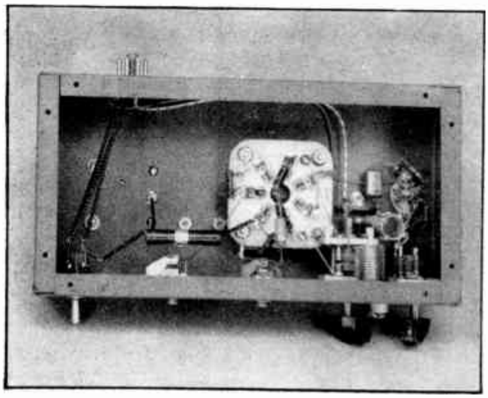

Fig. 17-18— Bottom view of the simplified v.h.f. exciter.

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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. 17-19, 17-20 and 17-21 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.

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Fig. 17-19 — 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.

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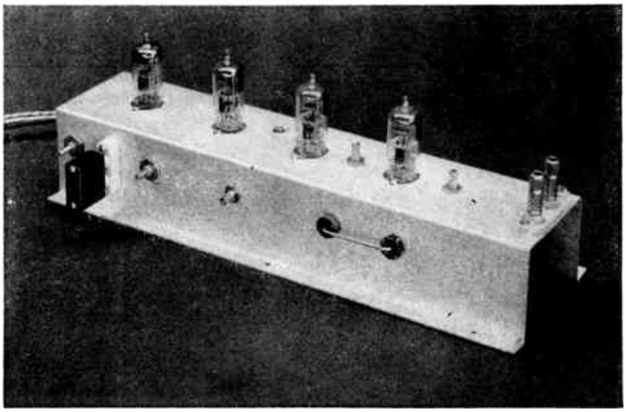

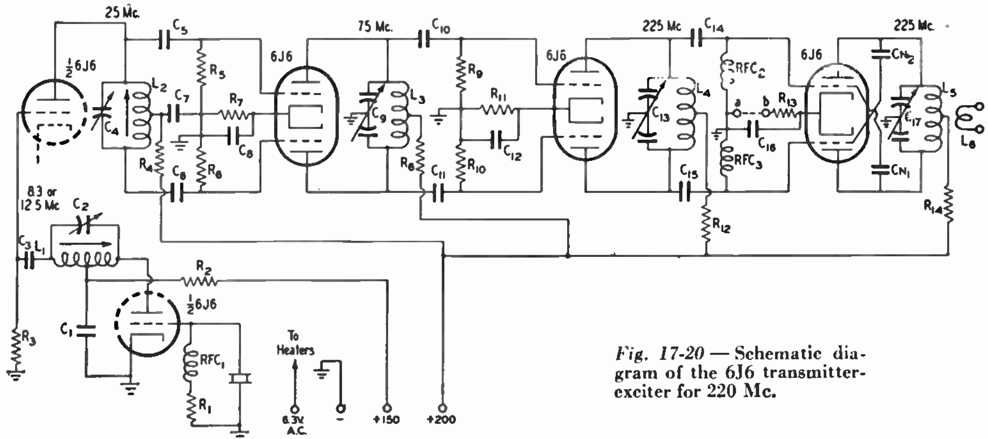

Fig. 17-20 — Schematic diagram of the 6J6 transmitter-exciter for 220 Mc.

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).
CN_1, CN_2 — 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. 24 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, 5/16 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. 17-17 is used to replace the more conventional crystal oscillator circuit of Fig. 17-20. 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. 17-20. It is suggested that the description of the 6- and 2-meter transmitter of Fig. 17-17 be studied carefully before this substitution is attempted.

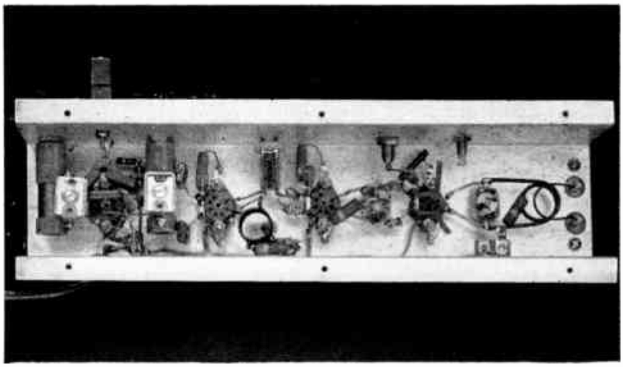

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Fig. 17-21 — Bottom view of the 6J6 220-Mc. rig, showing the simplicity of the layout.

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A Low-Powered Station for 50 and 144 Mc.

The two small transmitters shown in Fig. 17-22 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. 17-23) be closed, and S_1 in the 50-Mc. unit, Fig. 17-24, 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.

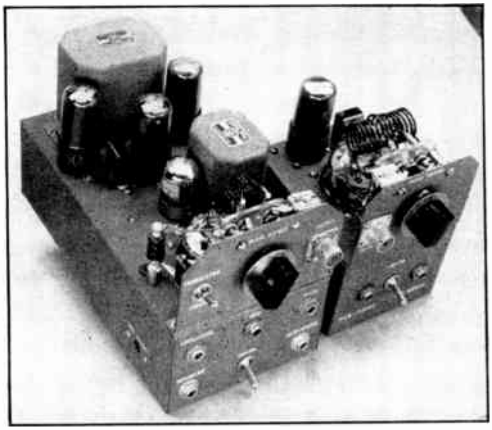

Fig. 17-22 — A 2-band set-up for mobile or low-powered fixed-station operation on 50 and 144 Mc. At the left is the 2-meter unit, complete with modulator. The smaller is the 50-Mc. r.f. section. Toggle switches permit use of the modulator with either r.f. section.

● THE 144-MC. SECTION

The 144-Mc. unit, Figs. 17-23 and 17-25, 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

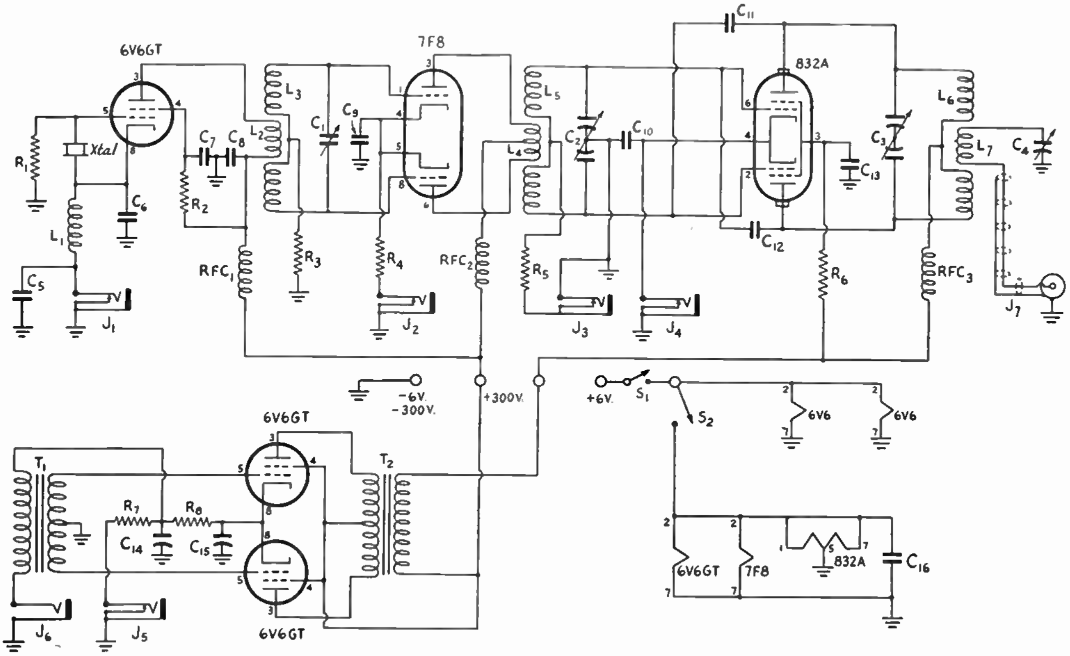

Fig. 17-23 — Circuit diagram of the 114-Mc. r.f. section and modulator.

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 standoff 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. 17-25). 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. 17-25 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.

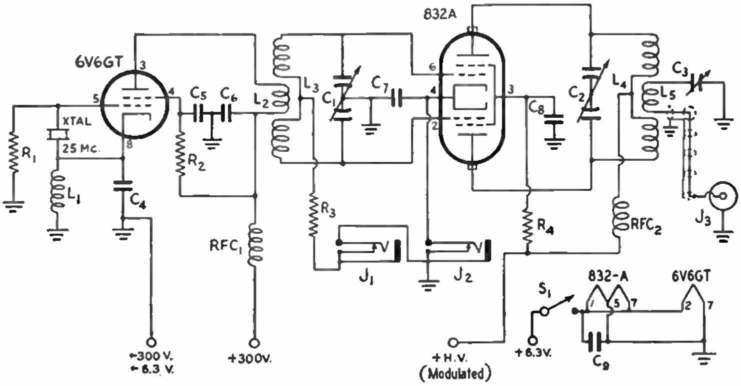

Fig. 17-24 — Circuit diagram of the 6-meter r.f. section.

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.

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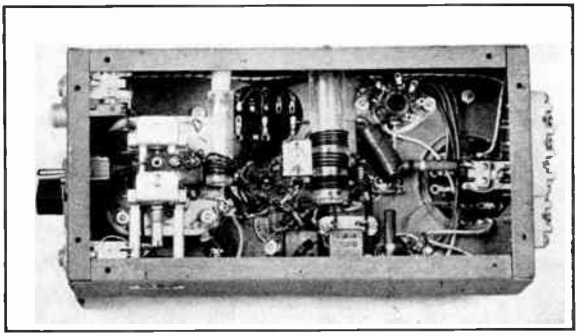

Fig. 17-25 — Bottom view of the 144-Mc. transmitter. The coil forms for L_2L_3 and L_4L_5 are mounted on the side wall of the chassis; the form for L_4L_5 is mounted on a small stand-off insulator so that the windings can be brought out to the center line of the chassis. C_1, the grid condenser for the 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.

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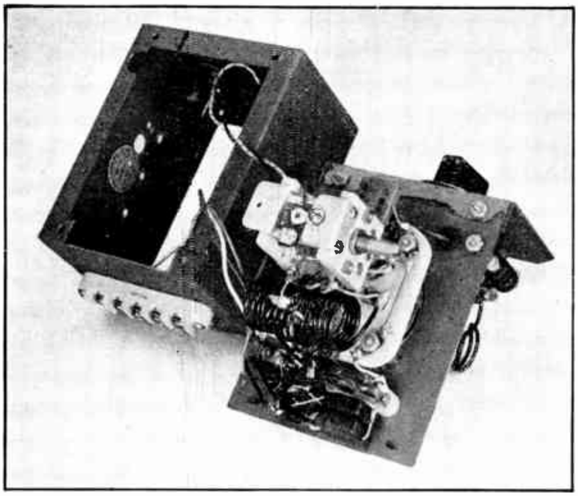

Fig. 17-26 — Bottom view of the mobile transmitter, showing all major components attached to the top plate.

THE 50-MC. PORTION

The 50-Mc. unit, shown in Figs. 17-24 and 17-24, 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. 17-22 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. 17-26. 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. 17-26) 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. 17-26, or a similar unit may be added, if only 50-Mc. operation is desired.

A 100-Watt R.F. Amplifier for 50 and 144 Mc.

The r.f. amplifier shown in Figs. 17-27, 17-28 and 17-29 is designed for use with a dual beam tetrode such as the 829B or AX-9903. It is capable of handling an input of up to 120 watts on c.w. or f.m. and about 100 watts on a.m. 'phone. The driver stage should have an output of 5 watts or more, to assure adequate driving power. The same general layout may be used with an 832A or 815, if a suitable value of grid resistor is used. The 815 also requires a different socket.

The amplifier is built on an aluminum chassis 3 by 4 by 17 inches in size, with practically all components mounted topside. The two-band tank circuit described on page 396 is used, to facilitate easy band changing and assure efficient operation on 144 Mc. Only the plate circuit is tuned. The grid coils are made to resonate with the input capacitance of the tube. The plate tuning condenser is cut down to a capacitance suitable for 144-Mc. used by removing plates, leaving two stator and three rotor plates in each section. The two stator plates left are those on either side of the stator connection lug. One rotor plate is removed from each end of the shaft and four from the middle.

The tube socket is mounted on a bracket 3⅝

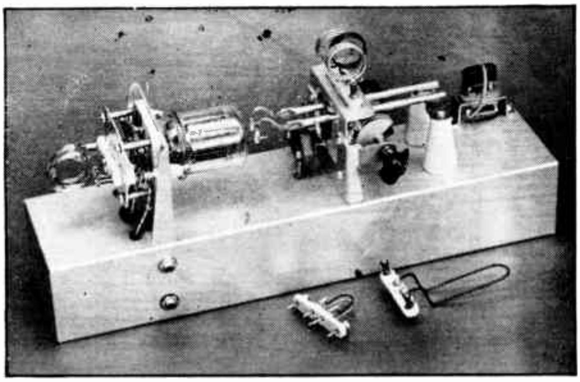

Fig. 17-27 — A dual-tetrode amplifier for 50 and 144 Mc., with 50-Mc. coils in place. In the foreground are the 144-Mc. grid coil and the antenna coupling loop used for 144-Mc. operation.

inches high, with the tube centered $2\frac{1}{2}$ inches above the chassis. The tuning condenser and coil socket are also mounted on brackets, the former $2\frac{3}{8}$ inches high. Both brackets have U-shaped cutouts to pass the plate lines with at least $\frac{5}{16}$ inch clearance all around.

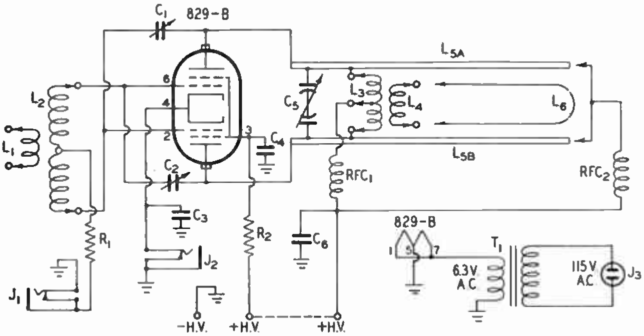

Fig. 17-28 — Schematic diagram of the two-band tetrode amplifier.

C_1, C_2 — Neutralizing capacitors, see text.
C_3, C_4 — 0.001-μfd. disc ceramic.
C_5 — Split-stator variable, approx. 15μμfd. per section (Millen 21935 with 2 stator and 3 rotor plates removed from each section).
C_6 — 0.001-μfd. mica, 1200-volt rating.
L_1 — 50 Mc.: 3 turns No. 18, $1\frac{1}{4}$-inch dia., turns spaced wire dia.
144 Mc.: U-shaped loop $\frac{1}{2}$ inch wide and $1\frac{1}{8}$ inch long, No. 11 tinned.
L_2 — 50 Mc.: 2 turns each side of L_1, same dia. and spacing, center tapped. Can be made by removing one turn from each end of a National AR-16 10-S assembly.
144 Mc.: U-shaped loop similar to L_1, but center tapped. See Fig. 17-27.
L_3 — 3 turns each side of center, No. 12 tinned, 1 inch dia., spaced 1 dia., center tapped. Leave $\frac{1}{2}$-inch space for L_4.
L_4 — 3 turns No. 14 enamel, 1-inch dia., spaced 1 dia.
L_{5a}, L_{5b} — $\frac{1}{4}$-inch o.d. copper tubing, $5\frac{1}{2}$ inches long, spaced $\frac{3}{4}$ inch on centers.
L_6 — Hairpin coupling loop $3\frac{1}{2}$ inches long, $\frac{3}{4}$ inch wide, No. 12 enamel.
J_1, J_2 — Closed-circuit jack.
J_3 — Male a.c. connector.
RFC_1 — 7.0-μh. r.f. choke (Ohmite Z-50).
RFC_2 — 1.8-μh. r.f. choke (Ohmite Z-144).
T_1 — Filament transformer, 6.3 volts, 3 amp.
R_1 — 4700 ohms, 1 watt.
R_2 — 10,000 ohms, 10 watts.

The plate lines are $5\frac{1}{2}$ inches long, exclusive of the flexible portion at the plate end. This is of tinned braid, making $1\frac{1}{4}$ inches additional, from the end of the lines to the slip-on connectors. The flexible portion of the line is made fast by inserting the end of the braid in the tubing and crimping the tubing in a vise. The connection is soldered for added firmness, but the tubing should be squeezed tight enough to hold the braid in place, as long periods of operation may heat the line sufficiently to loosen soldered connections. Connections from the lines to the tuning condenser are made by wrapping the tubing with four turns of tinned wire and soldering this wrap to the line and the condenser tab. The far end of the line is mounted on 2-inch standoffs and small copper brackets, bringing the over-all height to $2\frac{1}{2}$ inches.

The spacing of the lines, $\frac{3}{4}$ inch center to center, is determined by the spacing of the pins of the Millen 37212 plug used for a shorting bar. A short is placed across the terminals of the plug, and connection is made for the B-plus with a flexible lead. The Millen 37211 socket, mounted at the end of the chassis, serves as a convenient storage device for the plug and as a terminal strip for RFC_2. The plug may be used to adjust the line length; sliding it into or out of the line permits an adjustment of about $\frac{1}{4}$ inch in over-all length. This may be useful in counteracting for slight variations in tube characteristics.

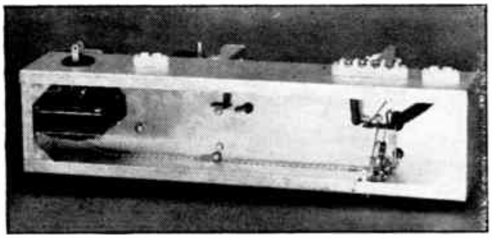

Fig. 17-29 — **Bottom view of the tetrode amplifier.**

The grid coil socket is mounted on a plate held in position by the screws on which the tube socket is mounted. It is positioned for minimum lead length — an important consideration. The input capacitance of the 829B is high enough so that it may be impossible to resonate the grid circuit at 148 Mc., if appreciable lead length or stray capacitance is introduced. If an 832A or AX-9903 is used the grid coil will be somewhat larger than that specified, and neutralization may not be needed.

Neutralization is accomplished, when required, by means of leads brought through the bracket, adjacent to the tube plates. These are crossed over to the opposite grids at the socket. Feed-through bushings are used and soldering lugs are attached to the bushings to provide the neutralizing capacitance. If more is needed these can be replaced with small tabs of sheet copper.

There may be a slight change in neutralizing capacitance needed for the two bands. As neutralization is inclined to be more critical at the higher frequency, the adjustment should be made carefully on 144 Mc. This same setting may be satisfactory for 50-Mc operation as well.

The plug-in coils are mounted on National PB-16 bases, fitting XB-16 sockets. When the stage is used on 144 Mc. the coupling is by means of a hairpin loop which plugs into the coil socket. The r.f. output is thus fed down to a crystal socket on the back of the chassis, for either band. A similar crystal socket is used for the r.f. input, at the tube end of the chassis.

Transmitting 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 and expense necessary to accomplish it. As the assignment is wide enough so that we need have little fear of interference problems there is no real need to employ advanced techniques in local work on 420 Mc., if one is not so inclined. Crystal control and selective receivers are desirable, but there is plenty of room for fellows who want to operate with simpler gear.

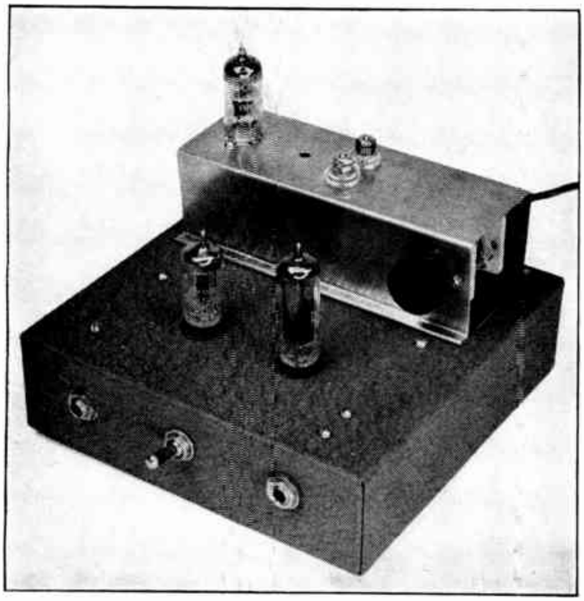

Fig. 17-30 — A 420-Mc. transmitter built in two units. The modulator portion, on a 7 × 7 × 2-inch chassis, uses a 6C4 driving a 6AQ5 modulator. The oscillator uses a 6J6 and is assembled on a removable trough-shaped chassis.

A SIMPLE LOW-POWERED TRANSMITTER

The transmitter shown in Figs. 17–30 through 17–33 is typical of the sort of thing that can be used to good advantage in developing local activity on 420 Mc. It runs only a few watts input, and delivers only about one watt of output, but it is quite capable of working over a radius of several miles when used with a good antenna system. A single 6J6 is used as a pushpull oscillator, with a halfwave line in its plate circuit. The complete oscillator assembly is built in a trough made of flashing copper. The 6AQ5 modulator and 6C4 speech amplifier are on the main chassis, at the back of which is a copper clip into which the oscillator unit is fitted. This arrangement permits experimenting with different types of r.f. sections without the necessity of making changes in the audio portion of the rig.

Only three adjustments are necessary in placing the unit into operation. The frequency should be checked with Lecher wires or a calibrated wavemeter, setting the frequency near the middle of the band. The method of determining the proper point for feeding the B-plus to the line is discussed earlier in this chapter. When this is done the coupling loop should be adjusted for maximum power in the antenna and the transmitter is ready for use. Frequency checks should be made again, after the antenna is connected to be sure that the signal radiated is well inside the band limits.

STABILIZED TRANSMITTERS

Not many presently-available transmitting tubes will operate satisfactorily on 420 Mc. Of the transmitting triodes only the 316A, 703A, 15E, 8012 and 8025 will work well as oscillators, and operation of these as frequency multipliers or amplifiers is generally not too satisfactory. Of the tetrodes the 832A and AX-9903 are the most useful for 420-Mc. work. Either will operate as a pushpull tripler to 420 Mc. with fair efficiency, giving enough output to drive a second tube of the same type as a straight-through amplifier. An 832A will give an output of about 2 watts as a tripler and 5 watts as an amplifier. The 9903 delivers 10 and 25 watts respectively. Both tubes should be operated at not more than about 60 percent of their normal ratings.

For best results tubes designed specifically for u.h.f. service should be used. These are the pencil-tube, lighthouse, and coaxial-electrode types, designed for use with coaxial or flat-plate tank circuits. The 5675, 2C43, 2C39 and 4X150 are typical of the special tubes that are capable of plate efficiencies up to 60 percent, when used in suitably-designed circuits. Forced-air cooling is necessary with these tubes when they are run at their normal ratings.

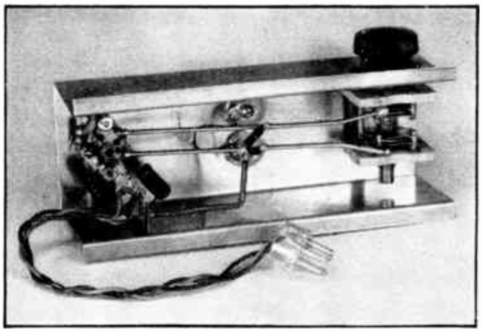

Fig. 17-31 — Bottom view of the oscillator assembly. The trough in which the components are mounted is made of flashing copper. It is 6 inches long, 1⅞ inches high, and 2¼ inches wide, with ¼-inch edges folded over for sliding into a clip attached to the main chassis.

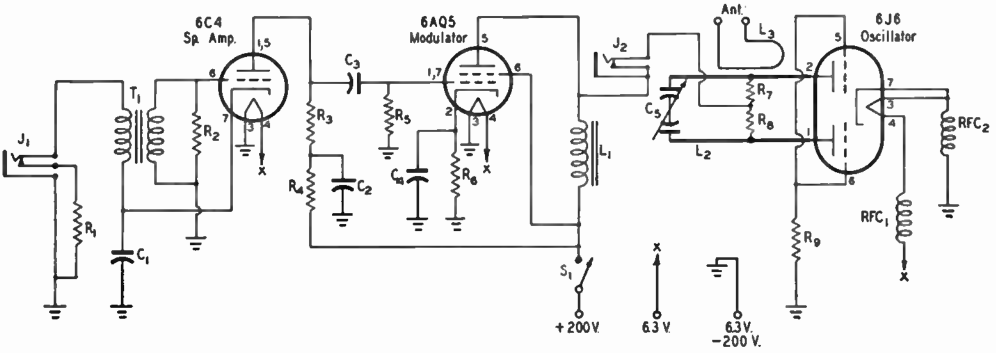

Fig. 17-32 — Schematic diagram of the 420-Mc. transmitter.

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 μμfd. per section. (Millen 21912D, with one rotor plate removed from each section.)
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, R_8 — 100 ohms, ½ watt, carbon.
R_9 — 2700 ohms, ½ watt.
L_1 — Midget filter choke.
L_2 — Plate line made of two pieces of No. 12 wire, 4¼ inches long, ⅜ inch apart, center to center.
L_3 — Hairpin of No. 18 wire. Portion which couples to L_2 is about ⅝ 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, 3/16-inch diam., ¾ inch long.
T_1 — Single-button microphone transformer.

Bibliography on 420-Mc. Equipment

For the convenience of the experimenter who is interested in 420 Mc. a list of articles appearing in *QST* 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.
"Better Results on 420 Mc." (Tilton), August 1950 *QST*, page 11.

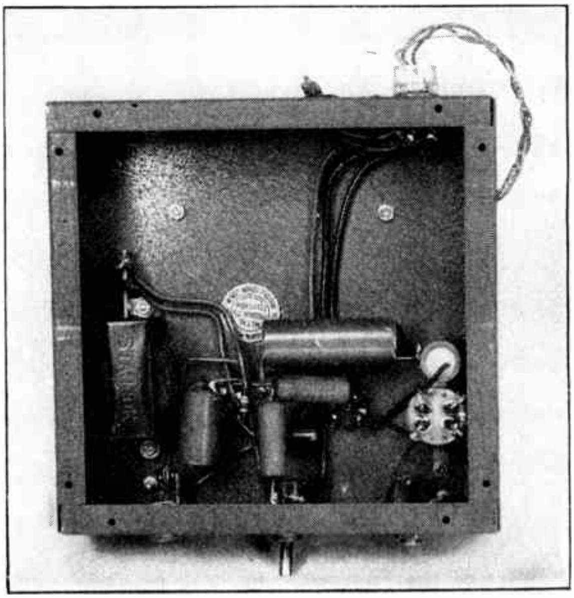

Fig. 17-33 — Bottom view of the main chassis of the 420-Mc. transmitter, showing audio components.

CHAPTER 18

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 may be 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

In the early days of v.h.f. operation everyone used simple antennas, and since the vertical half-wave gave at least as good performance in all directions as its horizontal counterpart offered in only two directions, the 5- and 2½-meter activity of the early '30s standardized on vertical polarization. Later on when high-gain antennas began to be used it was only natural that these, too, were put up vertical in areas where v.h.f. activity was already well established.

The increase in operating range that became available with the discovery of the various forms of long-distance propagation stirred interest in v.h.f. operation in areas where there was no previous experience, and here many of the newcomers started in with horizontal arrays, these having been more or less standard practice on the frequencies with which these operators were more familiar.

As best results are obtained only when the same polarization is used at both ends of a path, this use of both polarizations resulted in a conflict that, even now, has not been completely resolved. Numerous tests have demonstrated that there is little difference in results over long paths with either horizontal or vertical polarization, but each has features in its favor.

Horizontal systems are generally easier to build and rotate, particularly on 50 and 144 Mc. Where ignition and some other forms of man-made noise are troublesome, horizontal antennas usually provide better signal-to-noise ratio. Simple 3- or 4-element arrays are more effective horizontal than vertical, as their radiation patterns are broad in the plane of the elements and sharp in a plane perpendicular to them.

Vertical systems, on the other hand, can provide uniform coverage in all directions, a feature that is possible only with fairly complex horizontal arrays. Gain can be built up simply without introducing directivity, by stacking halfwave elements one above the other and feeding them in phase. This is a useful feature in net operation, or in locations where the installation of a rotatable system is not possible. Mobile operation is simpler and more effective when vertical antennas are used. The possibility of increased trouble with television interference, because of the use of horizontal polarization in TV broadcasting, has deterred v.h.f. men in densely populated areas from changing from vertical to horizontal.

The factors in favor of horizontal polarization have been predominant in 50-Mc. operation, and today we find practically complete standardization on horizontal systems for that band. The trend is toward horizontal for 144 Mc. and higher bands, particularly in long distance work, though vertical polarization is still widely used.

The newcomer to the v.h.f. bands should ascertain which is in use in the areas he expects to work, and go along with others in those areas. In setting up new activity in regions where no operation is presently going on, it is recommended that horizontal polarization be used, principally as a step toward eventual standardization.

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 400 to 600 ohms impedance, usually spaced about one to two inches; the polyethylene-insulated flexible lines, available in impedances of 300, 150, 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. 18-1. 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. 18-1. The "J" is also useful in mobile applications.

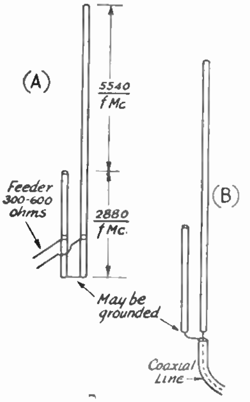

Fig. 18-1 — Two versions of the "J" antenna, often used in mobile installations, or in vertical arrays where parasitic elements may rotate around a fixed radiator.

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 the transmission-line chapter. 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. This consists of a quarter-wave line, usually of ¼ to ½-inch 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 the transmission-line chapter.

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 an arrangement shown in the transmission-line chapter, commonly called the "T"-match. It has the advantage of providing a means of adjustment (by sliding 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. Because the matching arrangement is adjustable, the dimensions of the "T" section are not critical. The position of the clips should be adjusted for lowest standing-wave ratio on the transmission line. The "T"-match may be used with any balanced line. This may be a double coaxial line or any two-wire system, either polyethylene insulated or open-wire construction. 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. 18-2. 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. 18-2 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. 18-4 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 monogram in the Transmission Lines Chapter. Note that center-to-center spacing of the conductors is important.

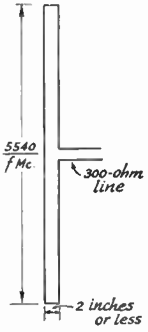

Fig. 18-2 — Details of the folded dipole.

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 another chapter, 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. 18-3. 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.

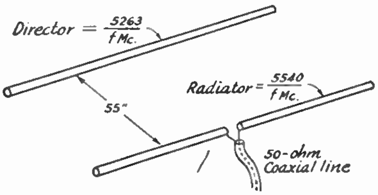

Fig. 18-3 — A simple 2-element array for 50 Mc. No matching devices are needed with this arrangement.

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 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 Fig. 18-4. 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.

The dimensions given are for peak performance at 50.5 Mc. For other frequencies, the

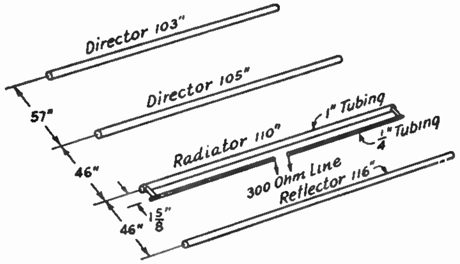

Fig. 18-4 — Dimensional drawing of a 4-element 50-Mc. array. Element length and spacing were derived experimentally for maximum forward gain at 50.5 Mc.

length of the folded dipole in inches should be figured according to the formula

$$L = \frac{5540}{f_{\text{Mc.}}}$$

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, slotted extensions may be inserted in the ends of the various elements, other than the dipole, as shown in Fig. 18-4. A 3-element array may be built, using the same general dimensions, except that the conductors of the folded dipole should have a 3-to-1 diameter ratio instead of 4-to-1.

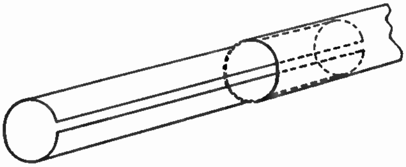

Fig. 18-4 — **Detail drawing of inserts which may be used in the ends of the elements of a parasitic array to permit accurate adjustment of element length.**

A Stacked Array for 50 Mc.

The radiation angle of a v.h.f. antenna system may be lowered, with a resulting improvement in operating range, by stacking two parasitic arrays one above the other and feeding them in phase. At spacings of ½ to ⅝ wavelength a gain of 4 db. or more may be realized by stacking in this way. An example is the 4-over-4 array for 50 Mc. shown in Fig. 18-5. The 12-element array mounted between the two 50-Mc. sections is described later.

TABLE 18-I

Dimensions for V.H.F. Arrays, in Inches

Freq. (Mc.)	50	144	220	420
Driven Element	110	38	24⅞	12¾
Reflector	116	40	26⅛	13⅜
1st Director	105	36	23⅝	12⅛
2nd Director	103	35¾	23⅜	12
Phasing Section*	114	39½	25⅞	13¼
0.25 Wavelength	57	19¾	13	6⅝
0.2 Wavelength	46	15¾	10⅜	5⅜
0.15 Wavelength	34	11¾	7¾	4

* Open-wire line only.

All-metal design is used in both arrays. 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 and held in place with U-shaped clamps of sheet aluminum, made similar to those in Fig. 18-9. Director spacing is 0.2 wavelength, reflector spacing 0.15 wavelength, see Table 18-I. 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.

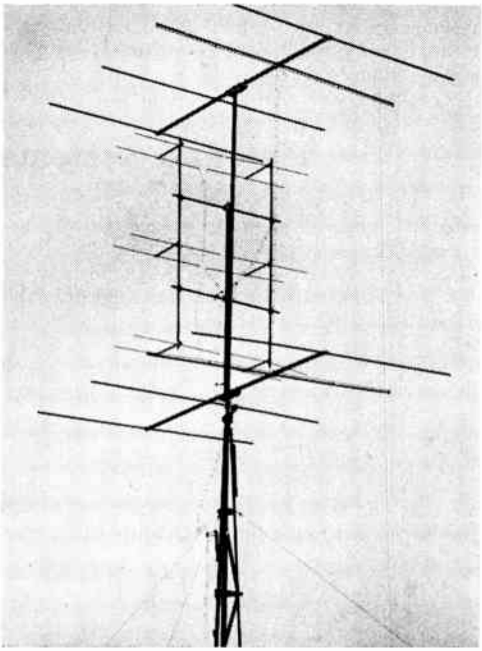

Fig. 18-5 — **An 8-element stacked array for 50 Mc., with a 12-element system for 144 Mc. mounted in the space between the two 50-Mc. sections.**

The main transmission line 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. No. 14 enameled wire spaced one inch is suggested. The feedline is brought at right angles from the phasing section to stand-off insulators on the main vertical support. It drops 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, it drops loosely to a fixed 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 $\frac{3}{16}$-inch copper tubing, mounted on $\frac{5}{8}$-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.

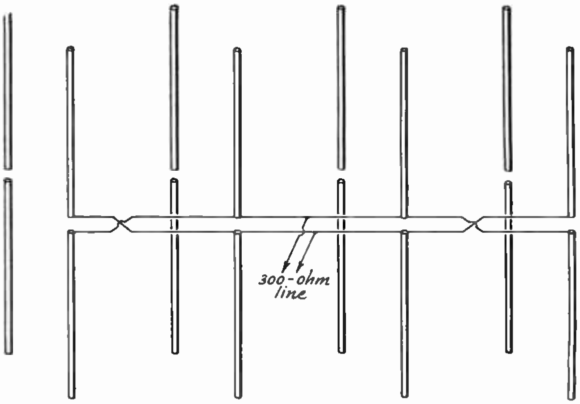

Fig. 18-7 — Schematic drawing of a 16-element array. A variable "Q" section may be inserted at the feed point if accurate matching is desired. Reflector spacing is 0.2 wavelength.

Phased Arrays for 144 Mc.

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. 18-5 to 18-7 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.

The 12-element array, Figs. 18-5 and 18-6, has a similar pattern in both horizontal and vertical planes. 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.

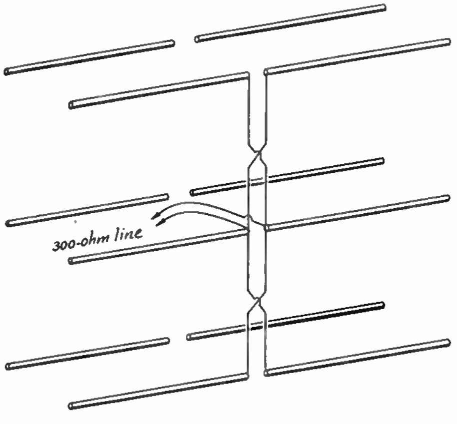

Fig. 18-6 — Element arrangement and feed system of the 12-element array. Reflectors are spaced 0.15 wavelength behind the driven elements.

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 $1\frac{1}{2}$ 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 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 18-I.

All-Metal Design in Phased Arrays

A very light weight yet physically rugged array can be built if the supporting structure is made entirely of metal. This involves few complications in the construction of small parasitic arrays, but with the 12- and 16-element jobs described above some special precautions must be observed. It is important that the supporting framework be designed so that it is entirely in back of the reflector plane, otherwise it will distort the radiation pattern and reduce the effectiveness of the system.

The supporting frame of the 12-element array of Fig. 18-5 is shown in Fig. 18-8, with a detail of the method of joining members in Fig. 18-9. When ¾-inch tubing is used for the frame, 1½-inch for the vertical support, and ¼-inch tubing for the elements, the sheet aluminum clamps may be made according to the dimensions given in Fig. 18-10. If other sizes of tubing are used the clamp dimensions can be determined readily by making experimental clamps of thin sheet metal or stiff cardboard. These can be folded and bent into the proper shape and then flattened out and used for templates.

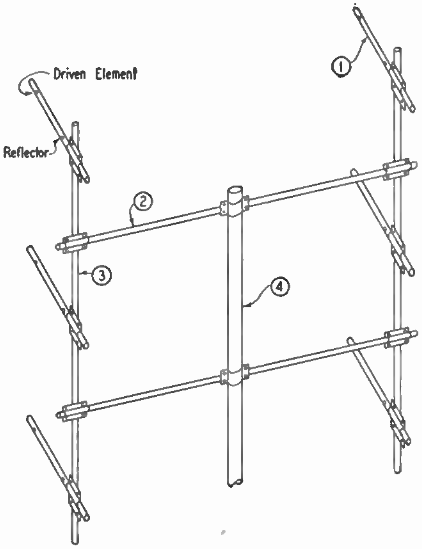

Fig. 18-8 — Supporting framework for a 12-element 144-Mc. array of all-metal design. Dimensions are as follows: element supports (*1*) ¾ by 16 inches; horizontal members (2) ¾ by 46 inches; vertical members (3) ¾ by 86 inches; vertical support (*4*) 1½-inch diameter, length as required; reflector-to-driven-element spacing 12 inches. Parts not shown in sketch: driven elements ¼ by 38 inches; reflectors ¼ by 40 inches; phasing lines No. 18 spaced 1 inch, 80 inches long, fanned out to 3½ inches at driven elements (transpose each half-wave section).

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 an earlier chapter. 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.

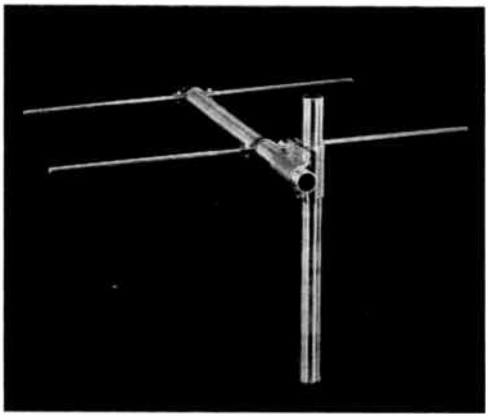

Fig. 18-9 — Model showing the method of assembling for all-metal construction of phased arrays. Dimensions of clamps are given in Fig. 18-10.

Matching devices that permit feeding long-wire antenna systems with flat lines also introduce one-band limitation, so their use is not advisable except in the case of 50 and 144 Mc., two bands that are close to third-harmonic relationship. A "Q" section that is approximately three quarter-wave-lengths long at 144 Mc. is one quarter-wavelength long at 50 Mc., so if the feed impedance of the antenna system is the same for both frequencies a "Q" section about 58 inches long may be used for both bands. In the case of a rhombic terminated in 800 ohms and fed with 300-ohm line, the matching section should have an impedance of about 500 ohms.

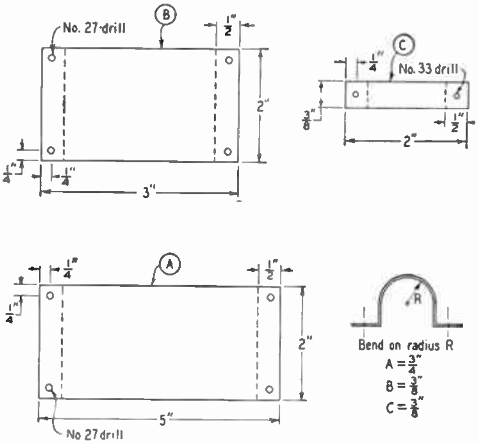

Fig. 18-10 — Detail drawings of the clamps used to assemble the all-metal 2-meter array. A, B and C are before bending into "U" shape. The right-angle bends should be made first, along the dotted lines as shown, then the plates may be bent around a piece of pipe of the proper diameter. Sheet stock should be 1⁄16-inch or heavier aluminum.

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. 18-6 and 18-7, may be adapted to use on 220 or 420 Mc. by using the dimensions given in Table 18-I.

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. 18-11. 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.

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 18-I.

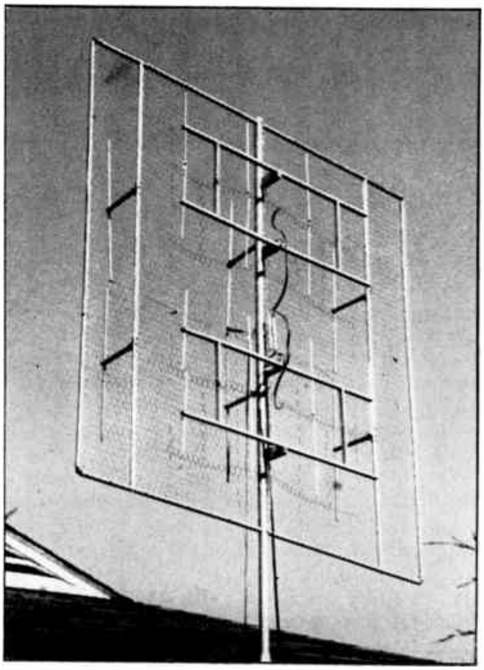

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Fig. 18-11 — A two-band screen-reflector array. One side has 16 driven elements for 420 Mc., and the reverse side has 6 half-waves in phase for 220 Mc. Both sets of elements are spaced 0.15 wavelength from the reflecting plane.

◆

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. 18-12 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 non-resonant, 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.

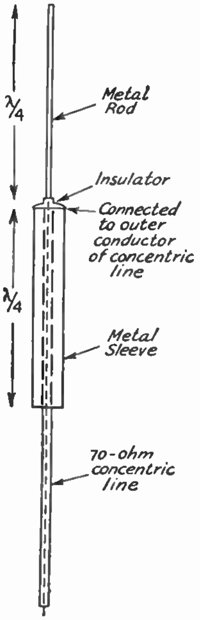

Fig. 18-12 — Coaxial antenna. The insulated inner conductor of the 70-ohm concentric line is connected to the quarter-wave metal rod which forms the upper half of the antenna.

Broadband Antennas

Certain types of antennas used in television work are of interest because they work across a wide band of frequencies with relatively uniform response. At very-high frequencies an 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 wideband antenna, on the other hand, will exhibit very nearly constant input impedance over several megacycles.

The simplest method of obtaining a broadband 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, spheroid, cone, diamond and double diamond. Of these, the conical antenna is perhaps the most interesting. With large angles of revolution, the variation in the characteristic impedance with changes in frequency can be reduced to a very low value, making such an antenna suitable for extremely wide-band operation. The cone may be made up either of sheet metal or of multiple wire spines.

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

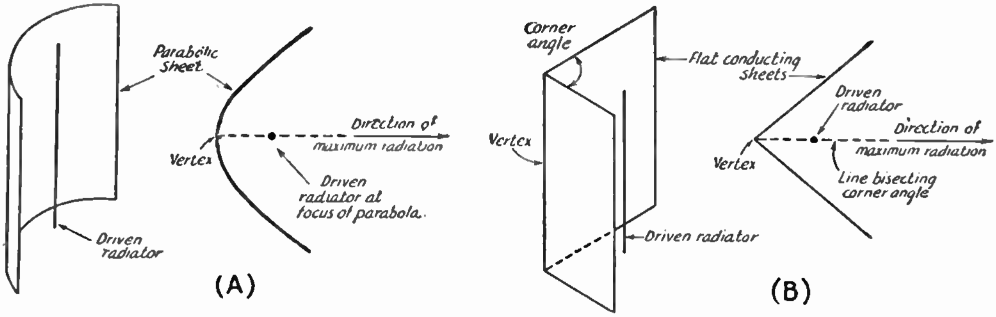

Fig. 18-13 — **Plane sheet reflectors for v.h.f. and u.h.f. A shows a parabolic sheet and B a square-corner reflector.**

wavelengths, sizes that may be practical for microwave work, 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. 18-13, 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.

CHAPTER 19

U.H.F. and Microwave Communication

In moving into the microwave region the amateur encounters marked differences in both the technical approach and the uses to which his frequency assignments may be put. Above 1000 Mc. we must discard most of our conventional circuitry and antenna ideas. Coils and condensers are replaced by resonant cavities. Parallel-wire transmission lines give way to coaxial lines or waveguide. Parasitic arrays are abandoned in favor of parabolic reflectors or horns. And in contrast to the random operating that has been so large a part of the amateur picture on our communication frequencies, microwave work is principally a matter of point-to-point communication between two cooperating stations.

These basic differences have tended to raise a natural boundary in the region around 500 Mc., beyond which relatively few communicating amateurs have ventured. The frequencies at the high end of the spectrum have a strong appeal to the experimenter, however, and new classes of licenses, now under discussion, are expected to provide the means whereby this type of worker may legally engage in two-way communication.

At least some amateur work has been done in all the assignments now open to our use. The work of these pioneers in adapting the frequencies above 1000 Mc. to communication purposes has been in line with the best amateur tradition, and it is hoped that the bands beginning at 1215 Mc. will see much amateur exploration in the near future. The frequencies assigned to amateurs in the microwave region are as follows: 1215 to 1300 Mc., 2300 to 2450 Mc., 3300 to 3500 Mc., 5650 to 5925 Mc., 10,000 to 10,500 Mc., and 21,000 to 22,000 Mc. Any frequency above 30,000 Mc. may be used. Any type of emission may be used in any of these bands, except in the case of the lowest, where pulse transmission is prohibited.

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 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. 19-1. 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.

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.

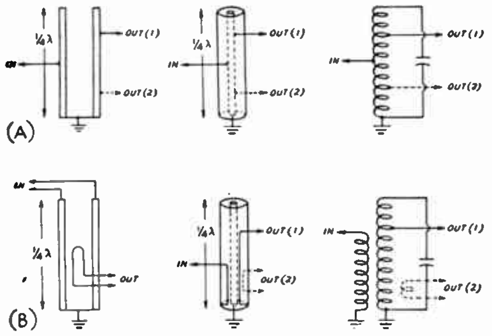

Fig. 19-1 — Equivalent coupling circuits for parallel-line, coaxial-line and conventional resonant circuits.

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. 19-2. 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.

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 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.

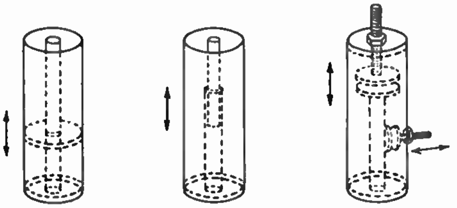

Fig. 19-2 — Methods of tuning coaxial resonant lines.

Two methods of tuning parallel-conductor lines are shown in Fig. 19-3. 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.

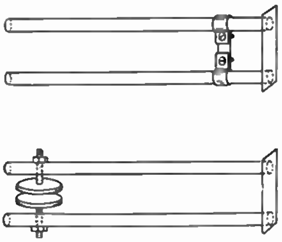

Fig. 19-3 — Methods of tuning parallel-type resonant lines.

Equivalent impedance points, for coupling or impedance-transformation purposes, are shown in Fig. 19-1 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.

"Butterfly" Circuits

The tank circuits described in the preceding section are primarily fixed-frequency devices. The "butterfly" circuits shown in Fig. 19-4 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 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.

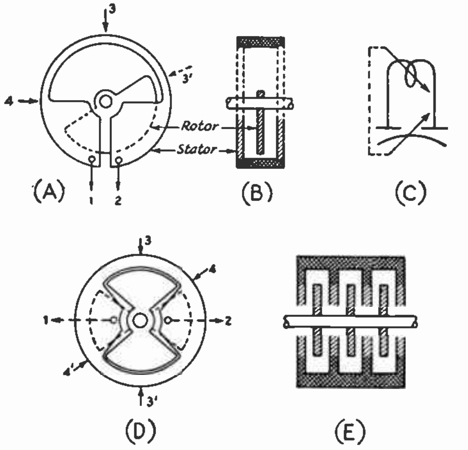

Fig. 19-4 — "Butterfly" tank circuits for v.h.f., showing front and cross-section views and the equivalent circuit.

The capacitance of any butterfly circuit may be computed by the standard formula for parallel-plate condensers given in the data chapter. 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.

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. 19-5A, 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 pipe shown in Fig. 19-5C. 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

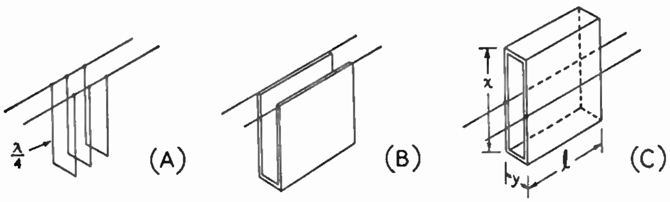

Fig. 19-5 — Evolution of a wave guide from a two-wire transmission line.

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.

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. 19-6. 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.

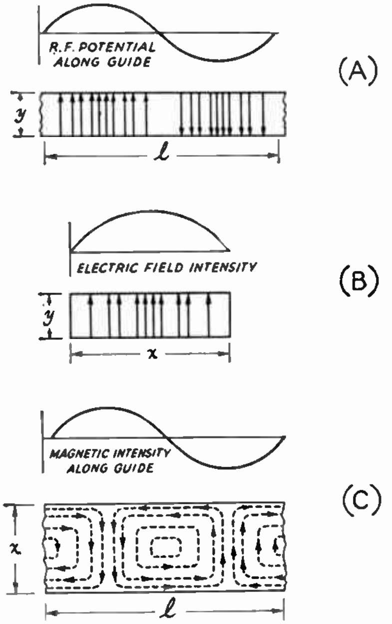

Fig. 19-6 — Field distribution in a rectangular wave guide. The $TE_{1,0}$ mode of propagation is depicted.

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. 19-7. 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. 19-7A 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.

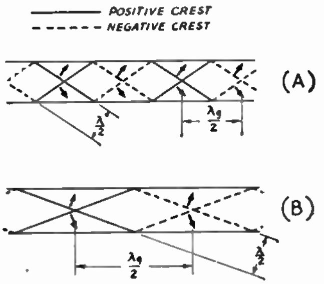

Fig. 19-7 — Reflection of two component waves in a rectangular guide. λ = wavelength in space, λg = wavelength in guide. Direction of wave motion is perpendicular to the wave front (crests) as shown by the arrows.

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. 19-7.

Modes of Propagation

Fig. 19-6 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 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. 19-5; 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. 19-8. 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. 19-8A 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. 19-8C, 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.

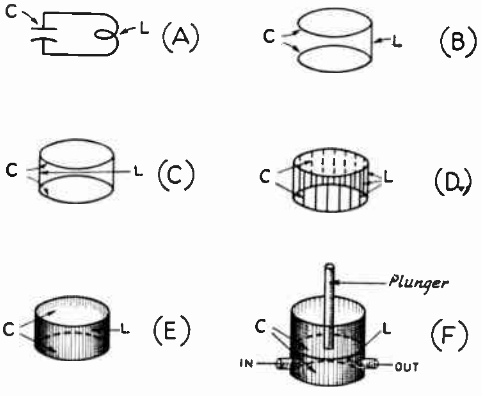

Fig. 19-8 — **Steps in the derivation of a cavity resonator from a conventional coil-and-condenser tuned circuit.**

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. 19-9. 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. 19-8F 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.

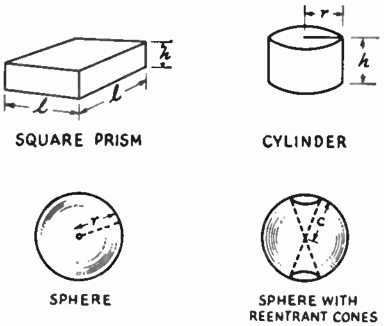

Fig. 19-9 — Forms of cavity resonators.

A form of cavity resonator in wide practical use is the re-entrant cylindrical type shown in Fig. 19-10. It is useful in connection with vacuum-tube oscillators of the types described for u.h.f. use elsewhere 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.

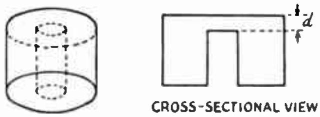

Fig. 19-10 — Re-entrant cylindrical cavity resonator.

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 are shown in Fig. 19-11. 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 electric 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.

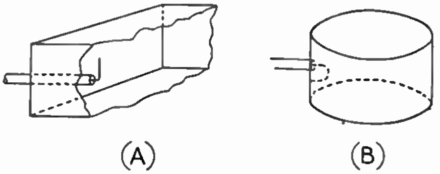

Fig. 19-11 — Coupling to wave guides and resonators.

U.H.F. and Microwave 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.

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. 19-12, 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 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.

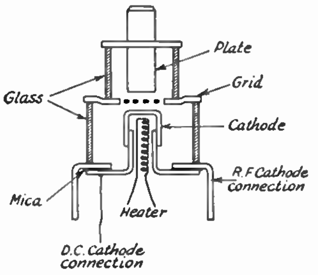

Fig. 19-12 — Sectional view of the "lighthouse" tube's construction. Close electrode spacing reduces transit time while the disk electrode connections reduce lead inductance.

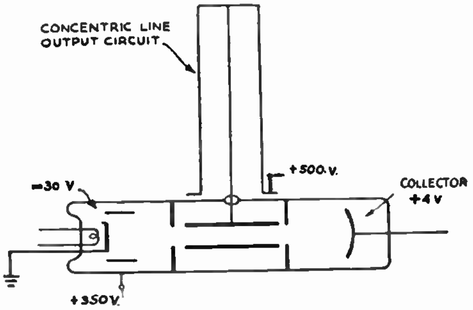

Fig. 19-13 — Simple form of cylindrical-grid velocity-modulated tube with retarding-field collector and coaxial-line output circuit, used as a superheterodyne high-frequency oscillator or as a superregenerative detector. Similar tubes can also be used as r.f. amplifiers and frequency converters in the 5–50-cm. region.

A simple form of velocity-modulation oscillator tube is shown in Fig. 19-13. 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.

If a feed-back loop is provided between the two rhumbatrons, as shown in Fig. 19-14, 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.

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.

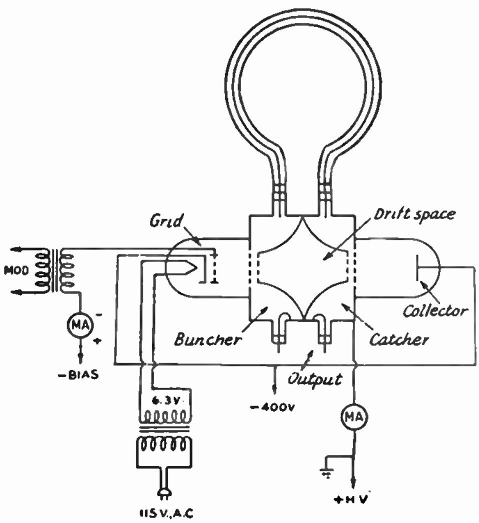

Fig. 19-14 — **Circuit diagram of the klystron oscillator, showing the feed-back loop coupling the frequency-controlling rhumbatrons and the output loop in the catcher.**

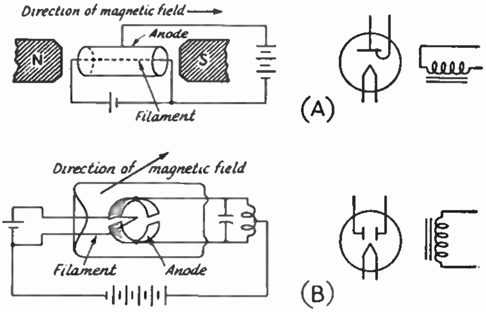

Fig. 19-15 — **Conventional magnetrons, with equivalent schematic symbols at the right. A, simple cylindrical magnetron. B, split-anode negative-resistance magnetron.**

In the transit-time magnetron the frequency is determined primarily by its dimensions and 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. 19-16. 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 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.

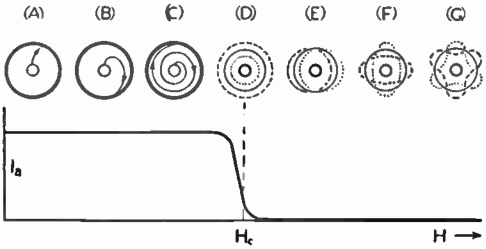

Fig. 19-16 — **Electron trajectories for increasing values of magnetic field strength, *H*. Below is shown the corresponding curve of plate current, I_a. Oscillations commence when *H* reaches a critical value, H_c; progressively higher-order modes of oscillation occur beyond this point.**

Fig. 19-16E, 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.

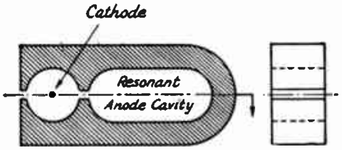

Fig. 19-11 — **Split-anode magnetron with integral resonant anode cavity for use at u.h.f.**

Split-anode magnetrons for u.h.f. are constructed with a cavity resonator built into the tube structure, as illustrated in Fig. 19-17. 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. 19-18.

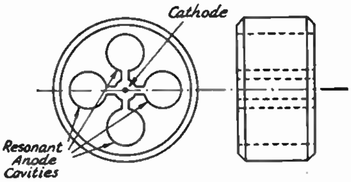

Fig. 19-18 — **Multisegment magnetron with four resonant cavities. This construction is used for extremely high frequencies.**

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.

Amateur Microwave Technique

All the bands that have been assigned to amateurs in the microwave region have been used for experimental two-way communication. Complete descriptions of suitable equipment for all these bands is beyond the scope of this text, but examples of the techniques employed are shown below. Reference is made to various articles that have appeared in *QST*, describing microwave gear used by amateurs, for those who wish more details.

1215 Mc.

In this band it is possible to use a few more-or-less conventional triodes with linear circuits, though great care must be used in designing such layouts, and the efficiency will be very low. A transmitter for 1215 Mc., designed and built by W3MLN and W3HFW, is shown in Figs. 19-19 — 19-21. It uses a 703A doorknob triode, completely shielded, with the antenna as an integral part of the assembly. The tube is mounted at the end of a halfwave line. Output is capacitively coupled to the folded quarter-wave antenna by means of a probe mounted alongside the plate line.

It should be emphasized that complete shielding of the oscillating circuit (including the tube elements) is absolutely necessary. The circuit will not oscillate at all if the shield is removed from the grid and plate rods, and only very weakly if

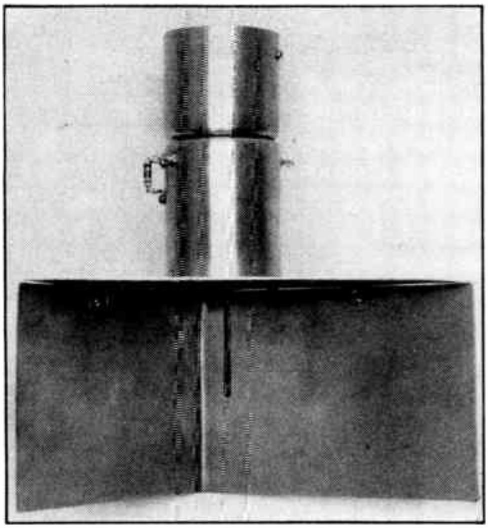

Fig. 19-19 — An oscillator and antenna system for 1215 Mc., built as one unit. (W3HFW — W3MLN)

the tube shield is not in place. Output is only about one watt, with an input of 80 ma. at 350 volts, but two of these units have been used to communicate over distances up to 12 miles or so with S9 signals. The equipment is described in detail by the designers in *QST* for April, 1948, Page 16.

Lighthouse tubes in suitably designed circuits are more efficient at this frequency. For best results cavities should be used, though trough-line and flat-plate circuits have been used.

Parabolic reflectors are usually employed for this and higher frequencies. It is desirable to make the transmitter or receiver an integral part of the antenna system if possible. If this cannot be done, coaxial line of the shortest usable length may be used. Air-insulated line is preferred to the flexible polyethylene-insulated variety, because of the higher losses in the latter.

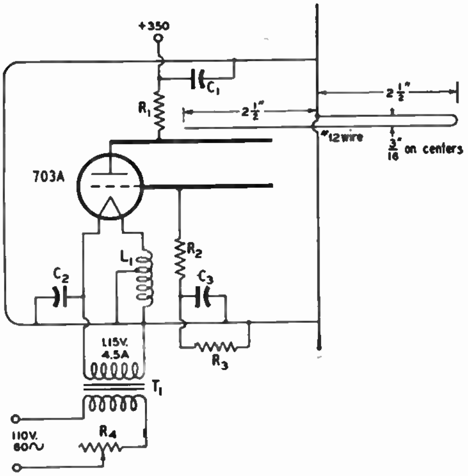

Fig. 19-20 — Schematic diagram of the 1215-Mc. oscillator.

2300 Mc.

Most of the work on 2300 Mc. has been done with lighthouse tubes in cavity oscillators, though some of the klystron types such as the 707B have been used. Cavities for this frequency may be a quarter wavelength, half wavelength or three-quarter wavelength long.

Details of a half-wave cavity oscillator using a 2C40 lighthouse tube are shown in Figs. 19-21 and 19-22. This oscillator was designed and built by W2RMA. It may be duplicated by any worker who has access to a few metal-working tools.

The main body of the cavity is 1-inch brass pipe, silver plated. The end that fits over the tube is cut out to an inside diameter of $1\frac{1}{32}$ inch, the only lathe work required. This end is also sawed crosswise at several points so that it may be clamped tightly to the tube with a brass strap, as seen in the photograph. Plate voltage is fed into the cavity through a feed-through capacitor mounted on the side of the tubing, and power is coupled out by means of a capacity probe and coaxial fitting at the hot end. The cavity is tuned with a screw mounted in the end, providing a variable capacitance to the anode post.

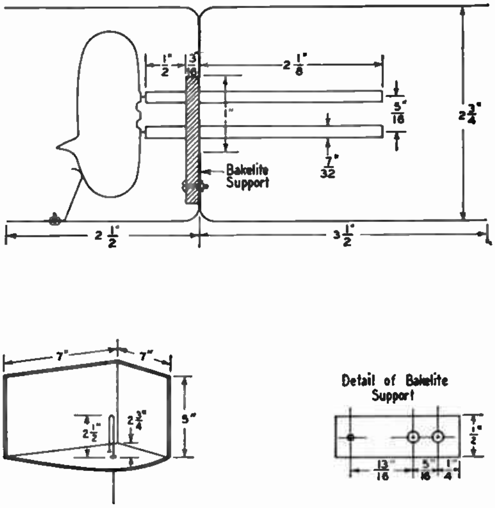

Fig. 19-21 — Detail drawing of the 703A oscillator for 1215 Mc.

Output, with a 250-volt supply, will be 50 to 250 milliwatts. This seemingly small amount of power may be made to do very well with the antenna gain that is possible at this frequency with a parabolic reflector of reasonable dimensions. Gear for 2300 Mc. is described in *QST* for July, 1946, Page 32, August, 1947, Page 128, and February, 1948, Page 11.

3300 Mc.

Lighthouse oscillators may be used on this frequency, but it is close to the top limit of their capabilities, so better results are obtainable with the klystron types. An advantage of the latter is that the frequency of oscillation may be varied over an appreciable range by changing the reflector voltage. This characteristic is also useful in providing a convenient means of obtaining frequency modulation. This sensitivity to voltage changes makes it desirable to use a regulated hum-free supply.

On this and higher frequencies a convenient system for two-way work is the use of a klystron as both transmitting oscillator and as a local oscillator for receiving. A crystal mixer is used in this case, its output being fed into a receiver serving as the i.f. system. If the receiver so used is capable of f.m. detection it is only necessary to modulate the klystron reflector voltage to provide f.m. communication of good quality. The oscillators of the two stations in communication are then operated on frequencies differing by the value of the intermediate frequency selected. A single antenna system is used for both transmitting and receiving, and no change-over arrangement is needed.

5650 Mc.

Amateur work in this range has been done largely with reflex klystrons, two types of which (2K43 and 2K44) are capable of operation within our band. The one-tube system described above may be used for each station, or of course separate tubes may be used for transmitter and local oscillator. In the latter case two antenna systems are required, but the transmitter efficiency is somewhat higher as some power is dissipated across the crystal in the one-tube arrangement.

Frequency modulation of klystrons is more practical than amplitude modulation. Modulation of the repellor voltage requires no audio power, as there is no current drawn by this tube element. A carbon microphone and a microphone transformer, with the repellor voltage fed through the secondary, will handle the audio requirements nicely.

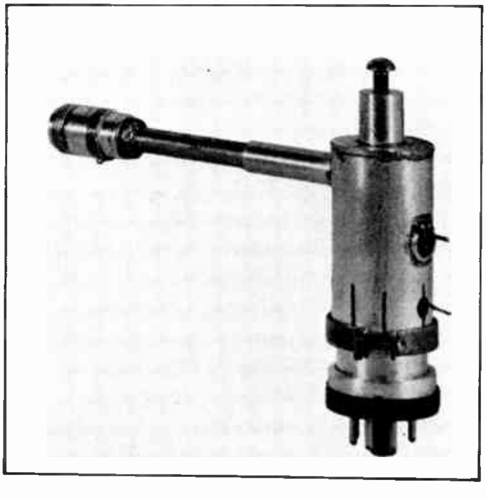

Fig. 19-22 — **A halfwave cavity oscillator for 2300 Mc. (W2RMA)**

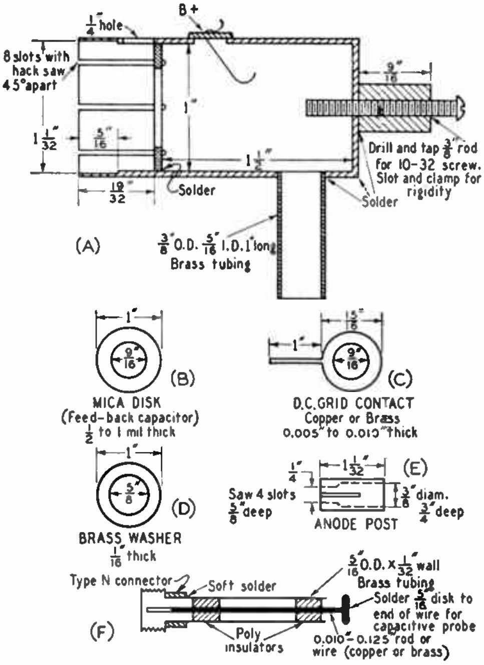

Fig. 19-23 — Mechanical details of the 2300-Mc. lighthouse oscillator.

The first two-way microwave communication in amateur history was carried out in this way by A. E. Harrison, W6BMS/2, and R. E. Merchant, W2LGF, who operated in the temporary 5300-Mc. band. Their equipment, described in *QST* for January, 1946, Page 19, will also work in the present band.

10,000 Mc.

The 723A/B reflex klystron, available at low cost for some time on the surplus market, provided amateurs with a convenient and inexpensive means of operation on 10,000 Mc. As manufactured, the tube will not ordinarily operate in the amateur band without modification.

Like other tubes of the reflex klystron variety, the frequency of oscillation is varied by warping the built-in cavity. It is used with a modified octal socket, with pin No. 4 removed and the hole enlarged to pass the coaxial line that is part of the tube. This line is terminated in an "antenna" which is ordinarily used to transfer power to a waveguide.

Two vertical struts are provided for tuning, one of which is already variable by means of a stud, which spreads or contracts the flexible strut on the right side, compressing or stretching the bellows, lowering or raising the frequency respectively.

The upper limit of frequency range, reached by rotating the tuning stud, will seldom be within the amateur band, hence it is necessary to perform the following operation. It may be seen that the top of the cavity is held in a fixed position on the strut on the side of the tube by two small nuts which, after having been tightened, have been spot-welded to each other. The spot weld should be filed away until each nut can be moved freely on the threaded stud. Next, the position of these nuts should be adjusted *very* carefully, to raise the top of the cavity as was done on the other side. Extreme care should be used in this operation, as excessive stretching of the bellows may break some of the seals and render the tube inoperative. It is advisable to move the lower nut only until a firm resistance is felt. The operating frequency should then be checked, and if it is still below the limit of the band another tube should be tried, as any further attempt to raise the frequency will almost certainly ruin the tube.

Equipment for use on 10,000 Mc. is described in detail in *QST* for February, 1947, Page 58.

21,000 Mc.

Operation in this frequency, and in the unassigned region above 30,000 Mc. is still highly experimental in nature. Only once has the 21,000-Mc. band been used for amateur two-way communication. This was accomplished under laboratory conditions by two engineers whose specialty is development work in this field. Their work is detailed in *QST* for August, 1946, Page 19. Type Z-668 reflex klystrons were used, with horn and parabolic antenna systems, to work two-way over a distance of 800 feet.

CHAPTER 20

Mobile Equipment

The amateur who goes in for mobile operation will find plenty of room for exercising his individuality and developing original ideas in equipment. Each installation has its special problems to be solved.

Most mobile receiving systems are designed around the use of a h.f. converter working into a standard car broadcast receiver tuned to 1500 kc. which serves as the i.f. and audio amplifiers. The car receiver is modified to take a noise limiter and provide power for the converter.

While a few mobile transmitters may run an input to the final amplifier as high as 100 watts, an input of about 30 watts normally is considered the practical limit unless the car is equipped with a special battery-charging system. The majority of mobile operators use 'phone.

In contemplating a mobile installation, the car should be studied carefully to determine the most suitable spots for mounting the equipment. Then the various unit should be built in a form that will make best use of that space. The location of the converter should have first consideration. It should be placed where the controls can be operated conveniently without distracting attention from the wheel. The following list suggests spots that may be found suitable, depending upon the individual car.

On top of the instrument panel
Attached to the steering post
Under the instrument panel
In a unit made to fit between the lower lip of the instrument panel and the floor at the center of the car
On the left-hand door panel (detachable when not in use)
Under the left-hand front seat
In the motor compartment (controls extended through the instrument panel)

The transmitter power control can be placed close to the receiver position, or included in the converter unit. This control normally operates relays, rather than to switch the power circuit directly. This permits a minimum length of heavy-current battery circuit. Frequency within any of the 'phone bands sometimes is changed remotely by means of a stepping-switch system that switches crystals. In most cases, however, it is necessary to stop the car to make the several changes required in changing bands.

Depending upon the size of the transmitter unit, one of the following places may be found convenient for mounting the transmitter:

In the glove compartment
Under the instrument panel
In a unit in combination with or without the converter, built to fit between the lower edge of the instrument panel and the floor at the center
Under the right-hand or left-hand front seat
On the ledge above the rear seat
Fastened to the back of the front seat
In the trunk
In the motor compartment

Most mobile antennas consist of a vertical whip with some system of adjustable loading for the lower frequencies. Power supplies are of the vibrator-transformer-rectifier or motor-generator type operating from the car storage battery.

Units intended for use in mobile installations should be assembled with greater than ordinary care, since they will be subject to considerable vibration. Soldered joints should be well made and wire wrap-arounds should be used to avoid dependence upon the solder for mechanical strength. Self-tapping screws should be used wherever feasible, otherwise lock-washers should be provided. Any shafts that are normally operated at a permanent or semi-permanent setting should be provided with shaft locks so they cannot jar out of adjustment. Where wires pass through metal, the holes should be fitted with rubber grommets to prevent chafing. Any cabling or wiring between units should be securely clamped in place where it cannot work loose to interfere with the operation of the car.

Noise Elimination

Electrical-noise interference to reception in a car may arise from several different sources. As examples, trouble may be experienced with ignition noise, generator and voltage-regulator hash, or wheel and tire static.

A noise limiter added to the car b.c. receiver will go far in reducing some types, especially ignition noise from passing cars as well as your own. But for the satisfactory reception of weaker signals, some investigation and treatment of the car's electrical system will be necessary.

Ignition Interference

The metal caps terminating ignition wires at the distributor usually are simply clamped onto the ends of the cables and thus depend upon an uncertain pressure contact with the wire. These wires should be fitted with new caps soldered to the conductor. The cable insulation should be inspected to make sure that there is no stray break-down between wires or to ground. Use fiber spacers to keep the cables away from ground and rerun them, if necessary, to keep them well spaced from low-voltage wiring that may carry noise through the firewall into the inside of the car.

The spark plugs should be kept clean and adjusted to proper gap. They, and the common distributor lead, should be fitted with good carbon suppressors. Before purchasing these resistors, it is a good idea to check them with an ohmmeter, since individual units may vary widely. A good resistor should measure within 20 per cent of 10,000 ohms. Sometimes r.f. chokes in series with the resistors will bring the noise down still further. Ohmite Z-28 chokes are usually quite effective in reducing noise on 10 meters. The motor timing should be readjusted after the insertion of suppressors. The distributor points should be in good condition, of course.

Generator Noise

Generator hash is caused by sparking at the commutator. It shows up as a high-pitched whine that varies with the speed of the motor. While the interference may not be noticeable in the b.c. band, it usually increases in intensity at the higher frequencies. At 4 Mc., and possibly 14 Mc., a large 15-volt electrolytic condenser (500 μfd. or more) connected between the generator output terminal and ground, alone or in conjunction with an r.f. choke in series with the output lead, may be sufficient. A few turns of No. 10 wire, space-wound, often will be enough. To reduce the noise at 28 Mc., it may be necessary to insert a parallel trap, tuned to the middle of the band, in series with the generator output lead. The coil should have about 8 turns of No. 10 wire, space-wound on a 1-inch diameter and should be shunted with a 30-μμfd. mica trimmer. It can be pretuned by putting it in the antenna lead to the home-station receiver tuned to the middle of the band, and adjusting the trap to the point of minimum noise. The tuning may need to be peaked up after installing in the car, since it is fairly critical.

Voltage-Regulator Interference

This type of interference may show up only at 10 and 11 meters. It is caused by sparking at the regulator points as they operate to reduce the charging rate when the battery approaches full charge. A condenser cannot be used across the contacts because it will cause the points to burn out in a short time. A satisfactory remedy for this type of noise is a toggle switch on the instrument panel that short-circuits the points when the switch is closed. This removes the interference and acts to provide full generator output. This does no harm so long as there is sufficient load on the battery to prevent overcharging. If a double-pole switch is used, it can be provided with a signal light to remind the operator to open the switch.

Wheel Static

Wheel static shows up as a steady popping in the receiver at speeds over about 15 m.p.h. on smooth streets. It is usually not noticeable on dirt, gravel or wet roads. It is caused by the grease in the front-wheel bearings insulating the wheels from the car. Front-wheel static collectors are available on the market to eliminate this variety of interference. They fit inside the dust cap and bear on the end of the axle, effectively grounding the wheel at all times. Those designated particularly for your car are preferable, since the universal type does not always fit well. They are designed to operate without lubrication and the end of the axle and dust cap should be cleaned of grease before the installation is made. These collectors require replacement about every 10,000 miles.

Rear-wheel collectors have a brush that bears against the inside of the brake drum. It may be necessary to order these from the factory through your dealer.

Tire Static

This sometimes sounds like a leaky power line and can be very troublesome even on the broadcast band. It can be remedied by injecting an antistatic powder into the inner tubes

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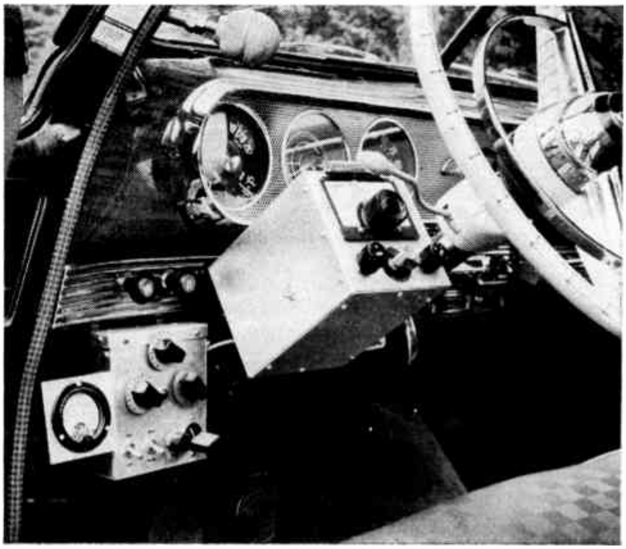

Fig. 20-1 — **The converter is clamped to the steering post, while the transmitter rides suspended from the instrument panel.**

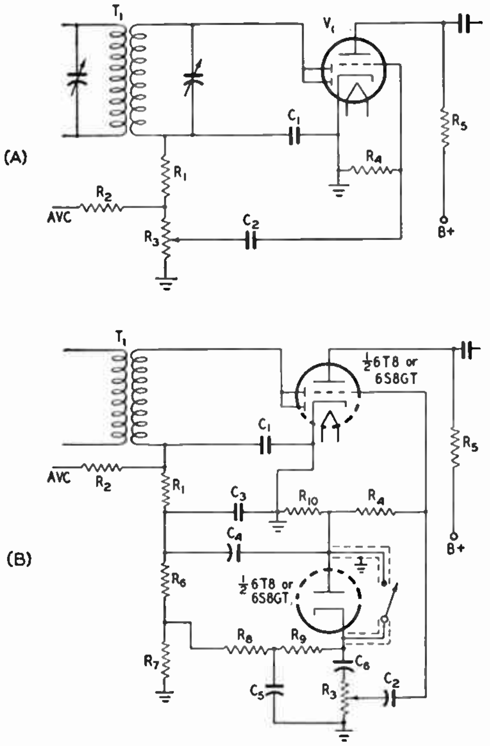

Fig. 20-2 — **Diagrams showing addition of noise limiter to car receiver. A — Usual circuit. B — Modification.**

C_1, C_3 — 100-μμfd. mica.
C_2, C_4, C_6 — 0.01-μfd. paper.
C_5 — 0.1-μfd. paper.
R_1 — 47,000 ohms.
R_2, R_{10} — 1 megohm.
R_3 — ½ megohm
R_7, R_8, R_9 — 0.47 megohm.
R_4 — 10 megohms.
R_5 — ¼ megohm.
R_6 — 0.1 megohm.
T_1 — I.f. transformer.
V_1 — Second detector.

through the valve stem. The powder is marketed by Chevrolet and possibly others. Chevrolet dealers can also supply a convenient injector for inserting the powder.

Tracing Noise

To determine if the receiving antenna is picking up all of the noise, the shielded lead-in should be disconnected at the point where it connects to the antenna. The motor should be started with the receiver gain control wide open. If no noise is heard, all noise is being picked up via the antenna. If the noise is still heard with the antenna disconnected, even though it may be reduced in strength, it indicates that some signal from the ignition system is being picked up by the antenna transmission line. The lead-in may not be sufficiently-well shielded, or the shield not properly grounded. Noise may also be picked up through the 6-volt circuit, although this does not normally happen if the receiver is provided with the usual r.f.-choke-and-by-pass-condenser filter.

With the motor running at idling speed or slightly faster, checks should be made to try to determine what is bringing the noise into the field of the antenna. It should be assumed that any control rod, metal tube, steering post, etc., passing from the motor compartment through an insulated bushing in the firewall will carry noise to a point where it can be radiated to the antenna. All of these should be bonded to the firewall with heavy wire or braid. Insulated wires can be stripped of r.f. by by-passing them to ground with 0.5-μfd. metal-case condensers. The following should not be overlooked: battery lead at the ammeter, gasoline gauge, ignition switch, headlight and taillight leads and the wiring of any accessories running from the motor compartment to the instrument panel or outside the car. For cars having Electrolok ignition systems, there is a special condenser that fits in the space in the top of the coil and by-passes the battery-supply wire from the ignition switch to the primary of the ignition coil. For other models, there is space in the top of the coil housing where a 0.02-μfd. 1000-volt mica condenser can be mounted. This measure is usually very effective, since it prevents the main source of noise from feeding into the interior of the car. The wire from the coil to the switch should be shielded, with the shield grounded to the firewall.

The firewall should be bonded to the frame of the car and also to the motor block with heavy braid. If the exhaust pipe and muffler are insulated from the frame by rubber mountings, they should likewise be grounded to the frame with flexible copper braid.

Noise Limiter

Fig. 20-2 shows the alterations that may be made in the existing car-receiver circuit to provide for a noise limiter. The usual diode-triode second detector is replaced with a type having an extra independent diode. If the car receiver uses octal-base tubes, a 6S8GT may be substituted. The 7X7 is a suitable replacement in receivers using loktal-type tubes, while the 6T8 may be used with miniatures.

The switch that cuts the limiter in and out of the circuit may be located for convenience on or near the converter panel. Regardless of its placement, however, the leads to the switch should be shielded to prevent hum pick-up.

A Bandswitching Mobile Converter

The circuit diagram of a bandswitching converter covering the 75-meter 'phone band and all of the 20-, 11- and 10-meter bands is shown in Fig. 20-3. To avoid the complication of tracking, the input circuit of the r.f. amplifier is tuned by a separate control. A single setting will hold over a considerable portion of each tuning range. The output circuit is broad-banded and thus requires only initial adjustment. By means of inductance slugs, it is tuned to the approximate center of each band. The high-frequency oscillator uses a high-C Colpitts circuit. Each of the bands is spread out over a good portion of the dial so there is no difficulty in tuning in and holding a signal. An air trimmer, C_9, is provided so that the tuning may be adjusted to calibration from the panel. The output coil, L_{14}, is tuned to 1500 kc. and is coupled to the input circuit of the b.c. receiver by L_{15}, a winding of several turns scramble-wound over the bottom end of the output coil, L_{14}.

A 5-circuit switch takes care of bandswitching in all circuits. One coil serves for both 27 and 29 Mc. at the input of the r.f. stage. A separate coil for 27 Mc. is required in the output circuit because a single coil does not quite cover both 27 and 29 Mc. satisfactorily. In the h.f. oscillator circuit, the same coil is used for both of the latter bands, but the tuning range is altered by switching in the series capacitance made up of C_{14} and C_{15}. C_{10} is added at 14 Mc. primarily for bandspread purposes, but it also improves the stability on this band.

One section of the bandswitch, S_{1E}, together with the final tap of S_{1A}, serves to connect the antenna to the b.c. receiver when the converter is not in use. The last switch section, S_{1F}, turns off the filaments of the converter automatically when the switch is turned to the b.c. position. Power for the converter is taken from an outlet added to the b.c. receiver. A dropping resistor in the b.c. set should be inserted if the "B" voltage to the converter exceeds 180 under load.

Construction

The converter shown in Fig. 20-4 is built on a 5 × 7 × 2-inch aluminum chassis. A box, 5⅛ inches high, made of sheet alumin-

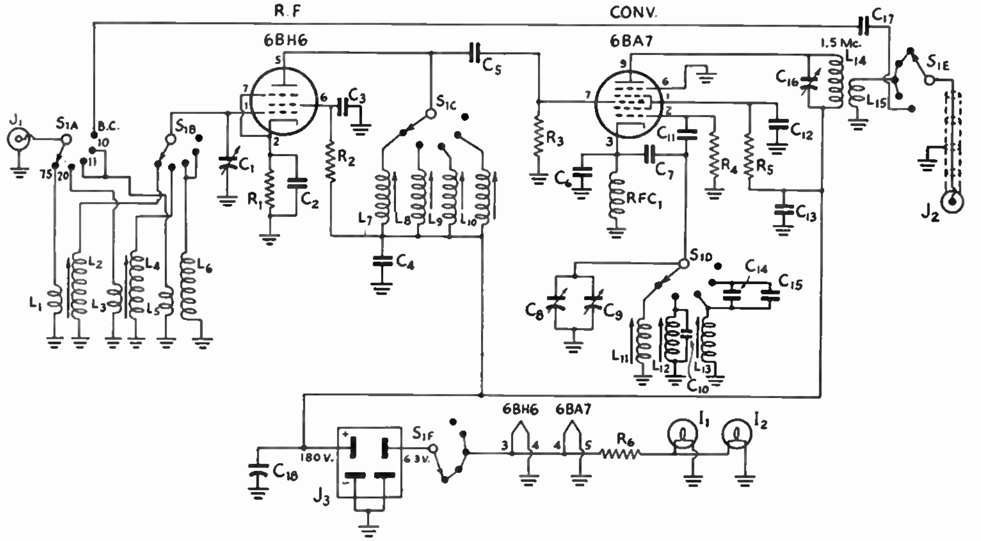

Fig. 20-3 — Circuit diagram of the mobile converter.

C_1 — 50-μμfd. variable air trimmer (National PSE-50).
C_2, C_3, C_4, C_{12}, C_{13} — 0.01-μfd. disc ceramic.
C_5 — 100-μfd. mica.
C_6, C_7 — 220-μμfd. silvered mica.
C_8 — Approx. 40-μμfd. variable (Millen 19050 with one rotor and one stator plate removed).
C_9 — Approx. 5-μμfd. variable (National PSE-25 with all but two plates removed).
C_{10} — 100-μμfd. silvered mica.
C_{11} — 47-μμfd. silvered mica.
C_{14} — 470-μμfd. silvered mica.
C_{15} — 330-μμfd. silvered mica.
C_{16} — 30-μμfd. mica trimmer.
C_{17} — 50-μμfd. ceramic.
C_{18} — 0.1-μfd. paper.
R_1 — 100 ohms, ½ watt.
R_2 — 22000 ohms, ½ watt.
R_3 — 15,000 ohms, ½ watt.
R_4 — 22,000 ohms, ½ watt.
R_5 — 15,000 ohms, 1 watt.
R_6 — 10 ohms, 1 watt.
L_1 — 15 turns No. 24 d.s.c. scramble-wound on L_2.
L_2 — CTC Type LSM-5 Mc., 7 turns removed. (40 μh.)
L_3 — 4 turns No. 24 d.s.c. wound at bottom of L_4.
L_4 — CTC Type LS3-10 Mc. (3.5 μh.)
L_5 — 3 turns No. 24 d.s.c. wound over bottom of L_6.
L_6 — 11 turns No. 22, ⅝ inch long, on CTC Type LS3 ⅜-inch diam. form, slug removed. (1 μh.)
L_7 — CTC LSM-1 Mc., 150 turns removed. (110 μh.)
L_8 — CTC Type LS3-5 Mc., 50 turns removed. (8 μh.)
L_9, L_{10} — CTC LS3-10 Mc., 4 turns removed. (1.9/2.2 μh.)
L_{11} — 90 turns No. 30 enam., wound on CTC Type LS4 ½-inch diam. form. (32 μh.)
L_{12} — 5 turns No. 20, ⅝-inch diam., spaced 16 turns per inch (B & W 3007 Miniductor), slipped over CTC Type LS4 ½-inch diam. form. (0.6 μh.)
L_{13} — 3 turns No. 16, ⅝-inch diam., spaced 8 turns per inch (B & W 3006 Miniductor), slipped over CTC Type LS4 ½-inch diam. form. (0.2 μh.)
L_{14} — CTC Type LS3-1 Mc., 80 turns removed. (450 μh.)
L_{15} — 20 turns No. 24 d.s.c. scramble-wound over bottom end of L_{14}.
I_1, I_2 — 6.3-volt 150-ma. dial lamp.
J_1 — Shielded jack (ICA 2378).
J_2 — Pin plug (ICA 2375).
J_3 — 4-contact chassis-mounting plug (Jones S-304-AB).
RFC_1 — 2.5-mh. r.f. choke (National R-50).
S_1 — Bandswitch — see text.

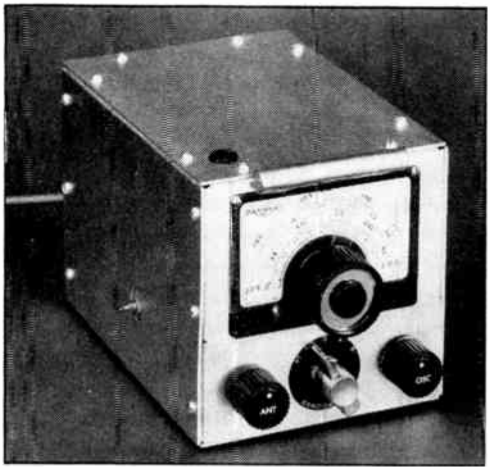

Fig. 20-4 — The completed bandswitching mobile converter ready to install. At the bottom, the r.f. input tuning is on the left and the oscillator trimmer on the right-hand side of the bandswitch.

um, is fitted around the chassis. Half-inch lips are bent over along the top and bottom edges of the sides, and along all four edges of the front and rear ends. The lips along the side edges of the front panel extend down only to the chassis. The box is assembled with machine screws and nuts. Four long machine screws through one side of the chassis provide means for attaching a clamp mounting so that the converter may be fastened to the steering post.

The National MCN dial is placed on the front panel so that it will line up with the shaft of the oscillator tuning condenser which is mounted directly on top of the chassis. It is necessary to notch out the front edge of the chassis for the dial mechanism.

The bandswitch is placed underneath at the center of the front edge of the chassis, with the controls for input tuning and oscillator trimmer to either side. The switch is made up from Centralab kit parts. All switch wafers are of the two-pole five-position type. One ceramic wafer (Type RR) is used for S_{1C} and S_{1D} in the amplifier-output and oscillator-circuits. This section is spaced two inches from the index head (Type P123). The other two wafers are of bakelite (Type K). The innermost of these serves for S_{1E} and S_{1F}, while the end section takes care of S_{1A} and S_{1B}. Two and one half inches back of the ceramic section, the two 6-inch switch-assembly rods pass through an aluminum bracket which provides a rugged brace for the rear end of the switch gang. The first bakelite switch wafer is spaced ¼ inch behind this bracket and the second wafer is ½ inch behind the first. The input tuning condenser, C_1, also is mounted on this aluminum bracket and is controlled by an extension shaft from the panel in front. The oscillator trimmer condenser, C_9, is fastened directly on the front edge of the chassis.

The placement of the two tubes can be seen in the top-view photograph of Fig. 20-5. The converter tube is near the front of the oscillator tuning condenser and the amplifier tube is to the rear, covered with a shield.

CTC (Cambridge Thermionic Corp.) slug-tuned coils and coil forms are used for the various inductances. Details are given under Fig. 20-3. About a half inch must be cut from the top of each of the LS4 forms so that they will fit under the chassis. The placement of the coils can be judged from the bottom-view photograph of Fig. 20-6. In that view, the oscillator coils are the three large ones near the bottom. From left to right, they are for the 75-, 20- and 10-11 meter bands. The three smaller coils above are in the output circuit of the r.f. amplifier. From left to right, they are for the 20-, 10- and 11-meter bands. The 75-meter coil is the large one above, mounted horizontally from the side of the chassis. The r.f. input coils are to the extreme left, grouped around the end of the switch. From top to bottom, they are for 10-11, 20 and 75 meters. The output coil, L_{14}, is hidden under the lip of the chassis in the extreme upper-right corner. Its tuning condenser, C_{16}, is the mica trimmer in the lower-right corner of the top-view picture (Fig. 20-5). A grommeted hole in the top cover permits adjusting this condenser after the top is in place. This may be found convenient in case it is necessary to shift the i.f. slightly to avoid interference from a strong local b.c. signal at 1500 kc. The inductance of L_7 is trimmed from the side, while the slugs of all other coils are adjusted before the cover is fastened down.

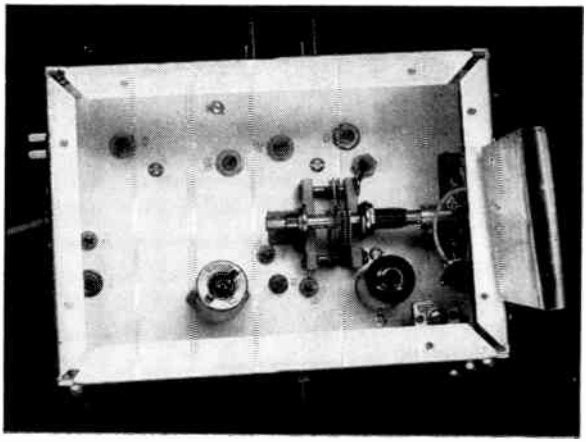

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Fig. 20-5 — Top view of the bandswitching converter with the cover removed to adjust the various inductance slugs. The object to the right is the dial-lamp shield.

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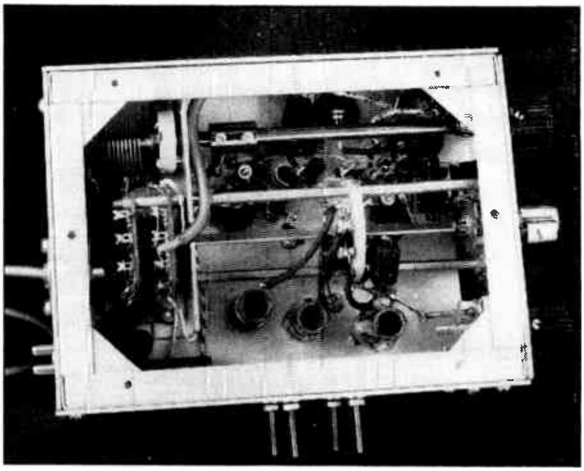

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Fig. 20-6 — Bottom view of the bandswitching mobile converter showing the placement of the miniature coils.

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A short length of coax line connects the output winding of L_{15} to the switch. Another external length connects the output of the converter to the input of the b.c. receiver. A pin jack at the rear provides a connection for the antenna input. Power connections are made at the rear through a four-contact connector.

Provision for illuminating the dial at night is made with a simple home-brewed arrangement. One end of a piece of shim brass or copper about 3 inches square is rolled a little more than halfway around a pair of standard 6.3-volt dial lamps placed butt to butt. The ends of the partial cylinder thus formed are covered by soldering in small discs of the same material. The lamps are then spaced about an inch apart and their shells are soldered to the metal enclosure. The two lead tips of the lamps are joined by a short piece of wire which connects to the "hot" side of the filament circuit. The remainder of the sheet is inserted between the top lip of the panel and the cover. By loosening the cover screws, it is possible to adjust the position of the lights for best illumination of the dial scale. The lamps should not need replacement often because the dimmer resistor, R_6, cuts the current down well below normal rating.

Adjustment

The output circuit of the converter tube should be adjusted first. Before proceeding, retrim the input circuit of the b.c. set to the antenna with the bandswitch in the b.c. position. Then switch to any of the four converter positions and tune C_{16} for maximum noise with the b.c. receiver tuning set at 1500 kc. The next step is to tune the h.f. oscillator to the appropriate ranges, starting with the 75-meter band. On all but the 10-11-meter band, the oscillator is tuned to the low-frequency side of the signal frequency. Since the i.f. is 1500 kc., the oscillator should be tuned 1500 kc. lower than the desired signal. For the range of 3800 to 4000 kc., the oscillator should cover the range of 2300 to 2500 kc. To accomplish this, turn the bandswitch to the 75-meter position, set the tuning condenser, C_8, at maximum and adjust the slug in L_{11} until the oscillator signal is heard on the station communications receiver at 2300 kc. (3800 minus 1500). To hear the signal, it may be necessary to run a wire from a point near the oscillator coil to the antenna terminal of the station receiver. Now, with an antenna connected to the input of the converter, swing the input tuning condenser, C_1, through its range, listening for a peak in noise. If none is found, set the slug in L_7 to a different position and try again. As soon as a noise peak is found on C_1, adjust the slug in L_7 for maximum response. The same procedure is followed for the 20-meter band, setting the tuning condenser at maximum, adjusting the slug in L_{12} until the oscillator is heard at 12,500 kc. (14,000 minus 1500), and then peaking up the r.f. stage input and output circuits. In this case, a second response point may be found. This is the image response to signals at 11,000 kc. If two response points are found, peak C_1 and L_8 at the response of higher frequency.

On the 10-meter band, which should be taken care of next, the oscillator is tuned to the high-frequency side of the desired signal. So, with the dial at the maximum-capacitance end, and the switch in the 28-Mc. position, adjust the slug in L_{13} until the oscillator signal is heard on the station receiver at 29,500 kc. (28,000 plus 1500), and then trim up the r.f. stage tuning as before. The image response will come at 31,000 kc., so be sure to peak up the r.f. circuits at the response of *lower* frequency.

Adjusting the slug in L_{13} for 28 Mc. also should place the oscillator in the correct range for the 11-meter band when the switch is in the 11-meter position. C_1 has sufficient range to cover both bands, but the separate r.f. stage output coil, L_9, must be peaked up. If it is found that the 11-meter range comes too far off on the dial, it may be necessary to slide the 10-meter range toward one end of the dial or the other by readjusting the slug in L_{12} slightly. As an alternative, the correction may be made by altering the capacitance of either series capacitor, C_{14} or C_{15}.

A Mobile Converter for 28 and 50 Mc.

The converter shown in Figs. 20-7 to 20-10 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. 20-8. 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. 20-7 — 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.* in Fig. 20-8. 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 intermediate 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. 20-9, 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, ½ 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 5⅛ inches wide. All three sides are 5 inches high with ½-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, 1⅞ 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 4⅞ by 1¾ inches in size, with flanges ½ inch wide folded over along the front and the bottom edges to provide a means of mounting. A 2¼ × 3¾-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 3½ inches wide and 4¾ 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

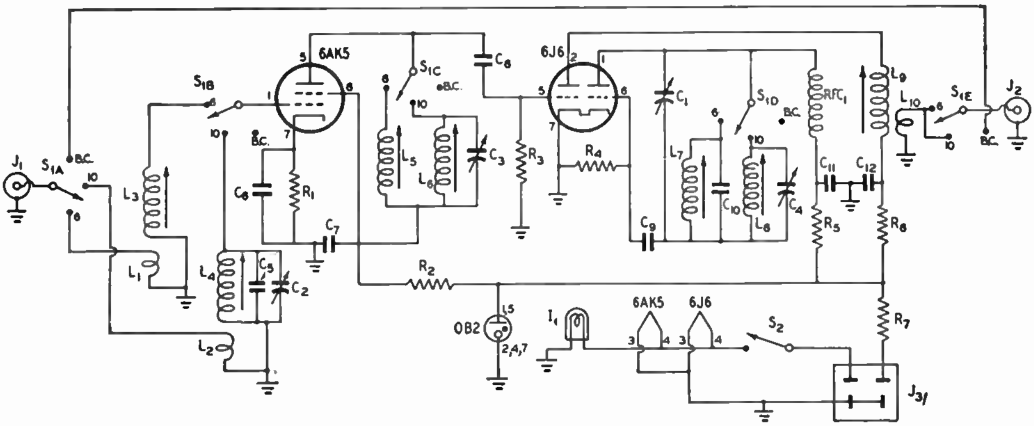

Fig. 20-8 — Circuit diagram of the bandswitching v.h.f. converter.

C_1 — 15-$\mu\mu$fd. variable reduced to one stator and 2 rotor plates (Millen 20015).
C_2, C_3, C_4 — 3–30-$\mu\mu$fd. mica trimmer (Millen 27030).
C_6, C_7 — 0.0015-μfd. ceramic (Centralab DA048002A).
C_8, C_9 — 100-$\mu\mu$fd. ceramic (Centralab CC32Z).
C_5, C_{10} — 10-$\mu\mu$fd. ceramic (Centralab CC20Z).
C_{11} — 500-$\mu\mu$fd. 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).
$S_{1\,A,\,B,\,C,\,D,\,E}$ — 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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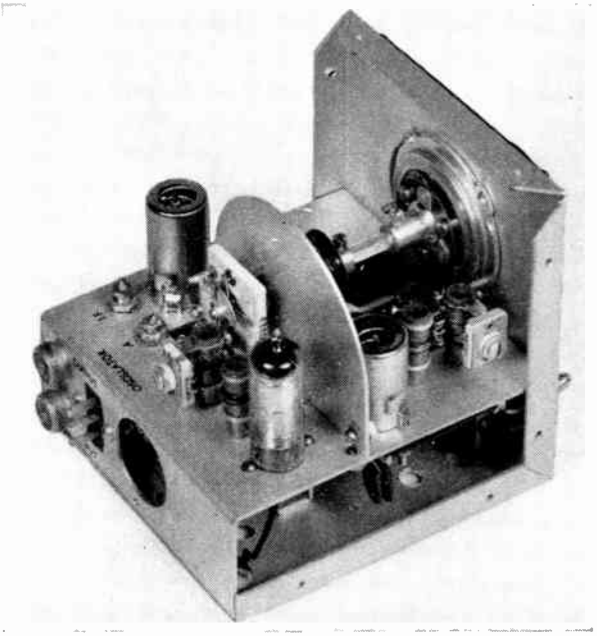

Fig. 20-9 — Interior view of the converter. Only the oscillator is tuned by the front-panel control, eliminating tracking problems.

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 0B2 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 photograph of the chassis, Fig. 20-10, shows how the circuit components are closely grouped around the tube sockets, with wiring completed to the point of making connections to the band-switch. Twin-Lead of the 75-ohm type is used to make the connection between the antenna input jack and the bandswitch. The two wires enclosed in spaghetti at the right of the chassis in the bottom view 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.

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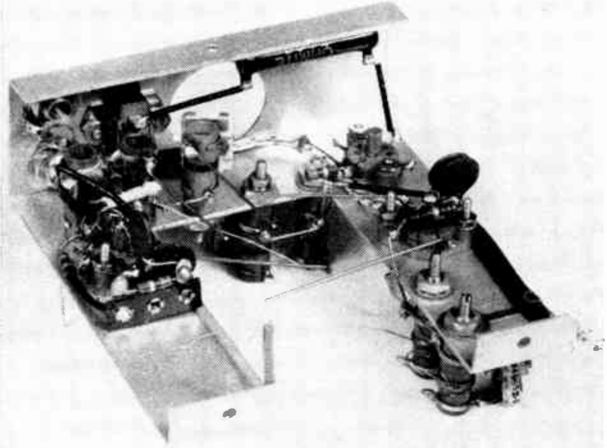

Fig. 20-10 — **Construction of the converter is made easier if as much wiring as possible is done before the assembling is completed. This bottom view of the chassis subassembly shows the wiring completed to the point of connection to the bandswitch.**

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A Mobile Converter for 144 Mc.

Working directly into the car broadcast receiver with a converter, as is done normally on lower-frequency bands, is not satisfactory for 144-Mc. work. Because the highest obtainable i.f. in such a system is 1600 kc., the image rejection is very low. Signals repeat within the band, making it very difficult to distinguish between the true signal and its image. A logical solution is to carry the conversion process one step further and design a 2-meter converter to work into a second converter designed for a lower frequency. The latter then feeds the signal into the car b.c. receiver.

This approach is employed by W1DBM in the 144-Mc. converter shown in the photographs of Figs. 20-11 through 20-14. The output frequency is 14 Mc., but it could just as well be 28 Mc. with suitable modification of the output coil, L_6. The output of this converter can be fed into the 28- and 50-Mc. converter described in the preceding section, or into any of the various manufactured converters on the market covering the 28- or 14-Mc. ranges.

Electrical and Mechanical Details

The circuit diagram, Fig. 20-13, shows a 6AK5 r.f. amplifier, 6J6 mixer-oscillator, and a 6AK5 i.f. amplifier. The pentode screens and the triode plates are fed through an OB2 voltage regulator. Slug-tuned coils are used in the 144-Mc. circuits for relatively broad response. The oscillator is tuned by a small split-stator condenser, with a ceramic padder for band-setting purposes. The mixer and i.f. plate coils are slug-tuned and are shunted with fixed ceramic condensers.

To hold down the overall size, some care must be used in planning the lay-out and assembly procedure. By mounting the tube sockets in the position shown, it is possible to use a single shield for both i.f. and r.f. stages. This shield (shown in the bottom-view photograph of Fig. 20-14) is notched to clear the tube prongs. The surrounding components must be mounted in such positions that the shield can be dropped into place and screwed down as a final operation.

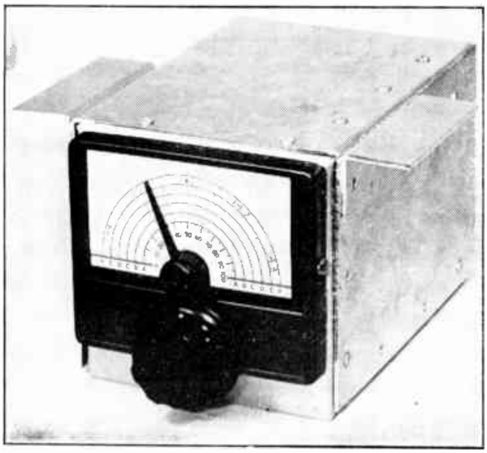

Fig. 20-11 — Mobile converter for 144 Mc. The heavy angle brackets are designed for mounting the converter under the dash.

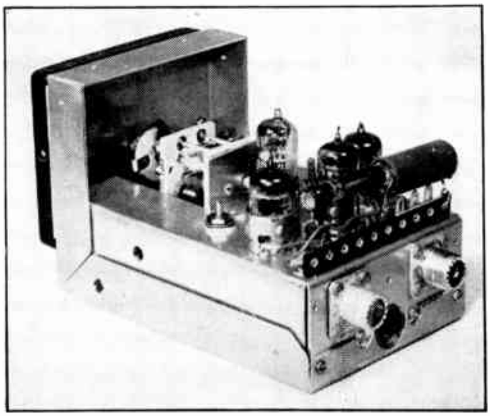

Fig. 20-12 — Rear view of the 2-meter mobile converter, with dust cover removed.

The panel is 4 by 4½ inches in size, this being determined by the dimensions of the Millen 10039 dial. The chassis is 5½ inches long, 4½ inches wide and 1¾ inches deep. With this depth, the OB2 regulator socket must be submounted, because it is of greater height than the 6AK5s and 6J6. The panel and bottom are folded from a single piece of 1/16-inch aluminum, with ⅝-inch lips turned up on the sides.

In addition to the holes for the tube sockets, the chassis has a cut-out for the tuning condenser. The condenser is mounted ruggedly on a heavy angle bracket in position so that its shaft lines up with the hole in the vernier dial. Coaxial connectors for the antenna and output, and a small shielded 3-wire receptacle are mounted on the rear edge of the chassis.

The dust cover is also of 1/16-inch aluminum sheet. It has a removable back plate, with clearance holes for the coaxial fittings and power plug. Two mounting angles of 3/32-inch aluminum are bolted to the top edges of the cover. These must be strong since they are used to fasten the converter under the instrument panel. The unit is completely wired and tested before mounting in the combination panel and bottom cover. A clearance hole in the side of the chassis provides for final adjustment of the oscillator padder.

Pretesting

The slug-tuned coils can be adjusted to approximately the correct settings by the use of a grid-dip meter. If the coils are made closely

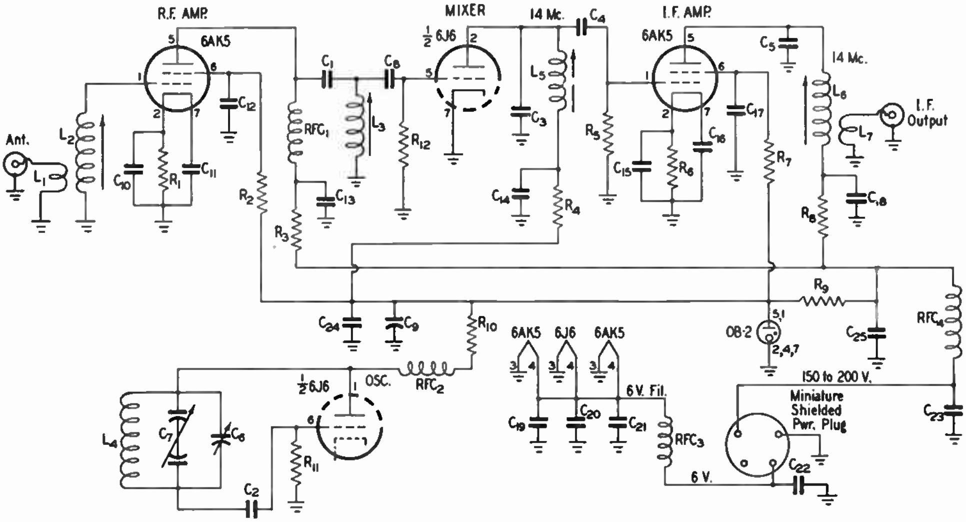

Fig. 20-13 — Wiring diagram of the 144-Mc. mobile converter.

C_1 — 3-μμfd. ceramic.
C_2, C_3, C_4 — 30-μμfd. ceramic.
C_5, C_8 — 50-μμfd. ceramic.
C_6 — 4–30-μμfd. ceramic padder.
C_7 — Miniature split stator, 2 rotor and 2 stator plates per section, double-spaced, double bearing.
C_9 — 4-μfd. 450-volt electrolytic.
C_{10}–C_{25} — 0.001- or 0.005-μfd. disc ceramic.
R_1, R_2, R_6, R_7, R_8 — 270 ohms, ½ watt.
R_3, R_4, R_{10} — 1000 ohms, ½ watt.
R_5 — 10,000 ohms, ½ watt.
R_9 — 10,000 ohms, 10 watts.
R_{11} — 15,000 ohms, ½ watt.
R_{12} — 1.5 megohms, ½ watt.
L_1 — 2 turns No. 20 enameled wire at cold end of L_2.
L_2 — 5 turns No. 20 enameled wire 5/16 inch long on CTC slug-tuned coil form ⅜-inch diameter, iron slug (approximately 0.17 μh.)
L_3 — 4 turns No. 20 enameled 5/16 inch long on CTC slug-tuned coil from ⅜-inch diameter, brass slug (approximately 0.08 μh.)
L_4 — 3 turns No. 12 tinned wire, ⅜ inch long, ⅜-inch inside diameter, with ¼-inch leads to condenser.
L_5, L_6 — 15 turns No. 28 enameled wire ¼ inch long on CTC slug-tuned ⅜-inch diameter coil form, combination iron and brass slug (approximately 2.5 μh.)
L_7 — 4 turns No. 28 enameled wire wound at cold end of coil form.
Values of L_5, L_6 and L_7 are for 14-Mc. i.f.
RFC_1, RFC_2, RFC_4 — 1-watt 1-megohm resistor wound full with No. 32 enameled wire.
RFC_3 — 1-watt 1-megohm resistor wound full with No. 18 enameled wire.

CTC coil forms (new ceramic type with high-frequency iron preferred) manufactured by the Cambridge Thermionic Corp., 546 Concord Ave., Cambridge, Mass.

to the dimensions given, it should be possible to adjust them and the oscillator padder, close enough to the proper values so as to be able to receive signals without further adjustments than these. The slugs should be adjusted before the heater and plate voltages are applied.

After this has been done, the converter should be placed in operation, using it in conjunction with the home-station communications receiver. The slug and padder settings should be rechecked, peaking the r.f. grid coil at 145 Mc., and the mixer grid coil at 147 Mc. for uniform response across the band. The mixer and i.f. amplifier plate coils can be peaked for maximum receiver noise. A slight readjustment of the oscillator padder condenser may be needed when the converter is installed in its case.

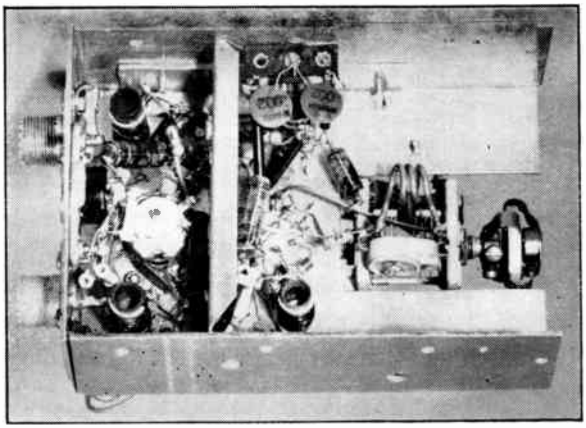

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Fig. 20-14 — Under-chassis view of the 2-meter converter. The coils at the bottom of the photo are (*left*) r.f. grid and (*right*) mixer grid. At the top, same order, are the i.f. amplifier and mixer plate coils. The latter is partially obscured by the small disc ceramics.

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A Bandswitching Transmitter for the Car

A complete compact bandswitching transmitter for mobile work, covering the 75-, 20-, 11- and 10-meter bands is shown in Fig. 20-15. The circuit diagram of the r.f. and control sections appears in Fig. 20-17. A 5763 miniature tube in the oscillator drives a 2E26 output stage. An octal tube socket, J_1, set in the front panel where it can be reached easily, serves as the crystal socket. With the crystal plugged into prongs 6 and 8, and prongs 4 and 7 shorted with a jumper, the circuit is a Tri-tet with the cathode tank adjusted for maximum 29-Mc. output from a 7-Mc. crystal. With this adjustment, adequate output is obtained also on the other bands. However, if desired, the circuit can be easily changed to a straight pentode circuit for straight-through operation at the crystal fundamental by plugging the crystal into prongs 4 and 6 and jumpering prongs 1, 4 and 8, thus cutting L_1 and C_1 out of the circuit. The octal socket also provides means for feeding a VFO, through prongs 6 and 8 with prongs 1 and 8 strapped together, to the input of the crystal stage. If desired, the extra socket prongs can be used to bring out power for the VFO, should one be used.

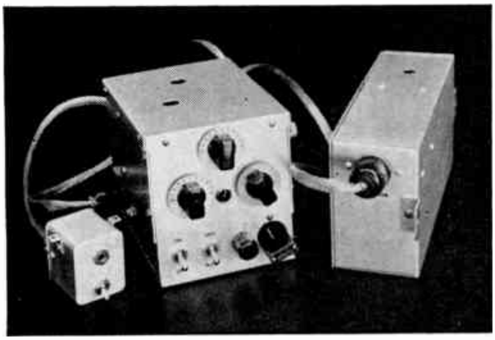

Fig. 20-15 — **A bandswitching mobile transmitter. From left to right are the control unit, the r.f. unit and the modulator.**

The oscillator stage is capacitively coupled to the amplifier. The plate of the latter stage is parallel fed. RFC_1 is a v.h.f. parasitic suppressor. A three-gang, three-position, rotary switch takes care of the bandswitching, changing coils in both stages and also the output coupling coils. The 10- and 11-meter bands are covered with a single set of coils.

Modulator connections are made at J_5. With a 5-wire cable connecting J_5 of Fig. 20-17 to J_1 of Fig. 20-20, the connections include power supply to the modulator. Connections between the r.f. unit and the power supply are made through J_6. This is a 5-prong plug to avoid any mistake in plugging in the cables. S_5 is the filament switch and it can be used also to control simultaneously power and antenna relays. A cable from a small control box plugs into J_3. The control-box wiring also is shown in Fig. 20-17. Connections to the microphone are included in the r.f. unit to avoid split cabling from the control box. S_3 is in the cathode circuit of the 2E26. This permits cutting off the amplifier while monitoring the oscillator which is still left running.

A milliammeter may be plugged in at J_4 to read either output-stage cathode current or grid current, depending upon the position of S_4. If desired, R_7 can be a multiplier shunt so that both currents can be measured with a single low-range (5- or 10-ma.) meter.

Modulator

The circuit of the modulator is shown in Fig. 20-20. It consists of a 12AT7 dual triode driving a 6N7 Class-B stage. The first section of the 12AT7 is operated as a grounded-grid amplifier so that a carbon microphone may be conveniently fed to the input without a transformer. Microphone voltage is obtained from the drop across R_1. A 3-circuit plug is needed, as indicated in Fig. 20-17. B_1 is a small 6-volt battery for bias.

Construction

In building the r.f. unit, an aluminum chassis, 5 × 7 × 2 inches is fitted with an aluminum box of sheet stock that makes the total height 6 inches. A strip 4½ inches wide is bent to fit around the sides and back of the chassis, overlapping the chassis by ½ inch along the bottom edges. The panel, which is 6 inches high and 5 inches wide, has ½-inch lips bent along the vertical sides down to the chassis level. These serve as a means for fastening the sides of the box to the panel. A partition 4 inches high and 5 inches wide, with ½-inch lips along the vertical edges, placed approximately 4 inches behind the panel, divides the enclosure into two sections. Approximately half-way up on the panel, the amplifier tuning condenser, C_{14}, and the bandswitch, S_1, are spaced with their shafts 1⅜ inches in from either edge. The antenna tuning condenser,

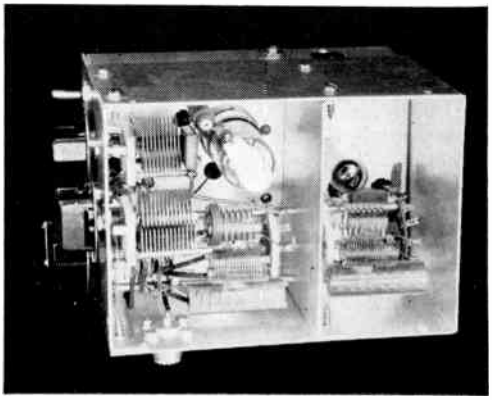

Fig. 20-16 — **Top view of the bandswitching mobile transmitter. The oscillator is at the rear and the amplifier and antenna-coupling section in front.**

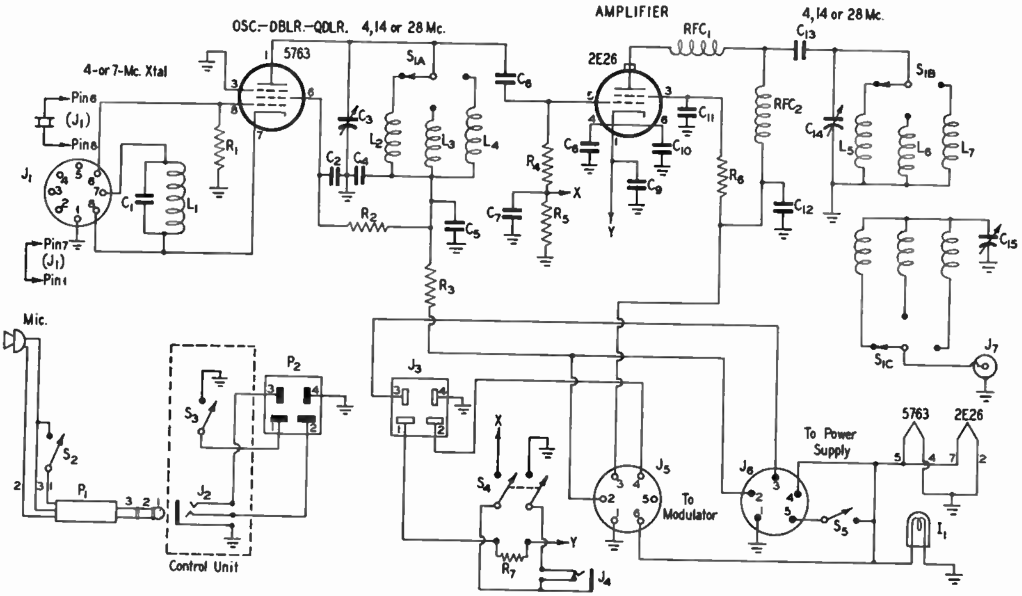

Fig. 20-17 — Circuit diagram of the bandswitching mobile transmitter.

C_1, C_6 — 100-μμfd. mica.
C_2, C_4, C_9 — 0.01-μfd. disk ceramic.
C_3 — 100-μμfd. variable (Millen 19100).
C_5, C_{12}, C_{13} — 0.002-μfd. mica.
C_7 — 0.005-μfd. disk ceramic.
C_8, C_{10} — 0.0001-μfd. disk ceramic.
C_{11} — 0.0068-μfd. mica.
C_{14} — 100-μμfd. variable (Millen 20100).
C_{15} — 140-μμfd. variable (Millen 20140).
R_1 — 0.1 megohm, ½ watt.
R_2 — 0.1 megohm, 1 watt.
R_3 — 1000 ohms, 10 watts.
R_4 — 18,000 ohms, 1 watt.
R_5 — 50 ohms, ½ watt.
R_6 — 20,000 ohms, 10 watts.
R_7 — 220 ohms, 1 watt.
L_1 — 2 μh. — 16 turns No. 22, ½ inch diam., ½ inch long (B & W 3004 Miniductor).
L_2 — 0.4 μh. — 7 turns No. 18, ⅝ inch diam., 1 inch long (B & W 3006 Miniductor)
L_3 — 1.2 μh. — 18 turns No. 18, ½ inch diam., 1 inch long (B & W 3003 Miniductor).
L_4 — 18 μh. — 64 turns No. 22, ⅝ inch diam., 2 inches long (B & W 3008 Miniductor).
L_5 — 0.5 μh. — 8 turns No. 18, ⅝ inch diam., 1 inch long (B & W Miniductor 3006); link 3 turns wound on outside.
L_6 — 2.4 μh. — 19 turns No. 22, ⅝ inch diam., 1¼ inches long (B & W 3007 Miniductor); link 4 turns of ½-inch Miniductor inserted.
L_7 — Same as L4; link 10 turns ½-inch Miniductor inserted.
I_1 — 6.3-volt signal lamp.
J_1 — Octal tube socket.
J_2 — 3-circuit microphone jack.
J_3 — 4-prong female connector (Jones S-404-AB).
J_4 — Closed-circuit jack.
J_5 — 6-prong tube socket.
J_6 — 5-pin chassis-mounting plug.
J_7 — Coaxial connector (Amphenol Type 1R).
P_1 — 3-circuit microphone plug.
P_2 — 4-prong plug connector (Jones P-404-CCT).
S_1 — Bandswitch — shown in 4-Mc. position (see text).
S_2 — Microphone-relay switch (included in microphone).
S_3 — Amplifier cathode switch — s.p.s.t. toggle.
S_4 — Meter switch — d.p.d.t. toggle.
S_5 — Filament switch — s.p.s.t. toggle.

C_{15}, is spaced centrally above. The indicator lamp, I_1, is mounted at the center. Along the bottom edge of the panel, from left to right in Fig. 20-15, are the filament switch, S_5, the meter switch, S_4, the control for the oscillator tuning condenser, C_3, and the crystal socket. Corresponding holes must be cut in the chassis. The output connector, J_7, is set in the right side of the box near the front corner.

After the panel has been drilled, the first part of the bandswitch should be assembled and mounted. The switch is made up of Centralab kit parts. It starts out at the front with the index assembly (Type P123). Section S1C (Type H) should be spaced ½ inch back of the index. A second similar switch section should be spaced ¼ inch back of the first. The contacts of this section are wired together; it serves only as a means of terminating the common ends of the output link coils. Spaced ¼ inch more toward the rear is a third switch section which also may be of bakelite (Type H). The contacts of this section also are wired together and the section serves as a termination and support for the common ends of the output tank coils. Spaced 2 inches farther to the rear is the fourth switch section, S_{1B}. This should be a ceramic wafer (Type X).

When the coils have been soldered in place, the two variable condensers may be mounted. Then the panel can be held temporarily in place while the socket for the 2E26 is located in the remaining available space. This done, the partition is drilled to fit the switch-assembly rods, including a clearance hole for the shaft. The partition is spaced ⅜ inch from the

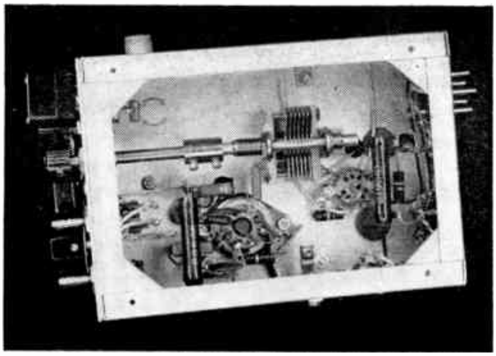

Fig. 20-18 — Bottom view of the bandswitching mobile transmitter.

last-assembled switch wafer. On the other side of the partition, spaced ½ inch, is another ceramic wafer, S_{1A}. The final section, spaced 1½ inches behind S_{1A} is of bakelite and this serves as a support and termination for the common ends of the oscillator plate coils.

The miniature oscillator tube is placed so that it will not interfere with mounting C_3 underneath the chassis. RFC_2 is placed alongside the 2E26 and RFC_1 is suspended between the plate cap and the top of RFC_2. The control connector, J_3, and the meter jack, J_4, are set in the left-hand side of the chassis, toward the rear, while the modulator and power connectors, J_5 and J_6, are mounted in the back edge.

The modulator components are enclosed in a 5 × 10 × 3-inch aluminum chassis, as shown in Fig. 20-19. With the exception of the modulation transformer and the biasing battery, the parts are mounted on a shelf or partition that spans the chassis. An aluminum strap clamps the battery against one side. Half-inch holes near the tubes provide ventilation.

The control unit, included in Fig. 20-15 is enclosed in a National Type RO shield can whose depth has been reduced to a little over 2 inches. The open end is closed with a piece of aluminum cut to fit and fastened in with small angles. Tabs are provided for fastening to the lip along the lower edge of the instrument panel in the car. The interconnecting cables are shielded in copper braid.

The power-control relay winding connects across Pins 3 and 4 of J_6. The ungrounded side of the battery connects to Pin 5 and plus h.v. to Pin 2.

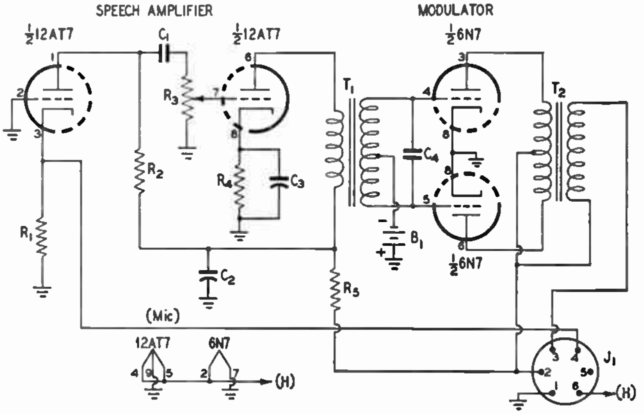

Fig. 20-20 — Circuit diagram of the modulator for the bandswitching mobile transmitter.

C_1 — 0.01-μfd. disk ceramic.
C_2 — 8-μfd. 450-volt electrolytic.
C_3 — 20-μfd. 25-volt electrolytic.
C_4 — 0.001-μfd. disk ceramic.
R_1 — 470 ohms, 1 watt.
R_2 — 22,000 ohms, 1 watt.
R_3 — 0.5-megohm potentiometer.
R_4 — 1500 ohms, ½ watt.
R_5 — 4700 ohms, 1 watt.
B_1 — 6-volt dry battery.
T_1 — Interstage transformer, 1:2 ration (Thordarson T20A16).
T_2 — Modulation transformer: 10,000 ohms plate to plate $5000 ohms (Stancor A3845).

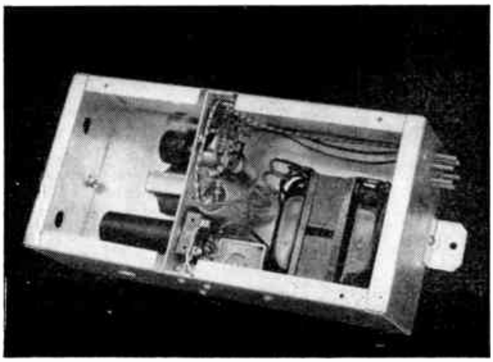

Fig. 20-19 — Inside view of the modulator for the bandswitching mobile transmitter.

Adjustment

The transmitter will operate satisfactorily with any supply delivering 300 to 500 volts. Naturally, the power output will be commensurate. With a 300-volt supply, the total current, including that of the modulator, and with the output stage loaded to about 55 ma., will run approximately 90 ma. and up to 125 ma. with modulation. At 450 volts, the total current will be near 130 ma. increasing to 170 ma. under modulation, the output-stage cathode current running about 130 ma. fully loaded. The output stage should operate properly with a grid current of 0.5 to 2 ma. If it exceeds this value the output circuit of the oscillator should be detuned to limit the grid current. The audio control should be preset to give adequate modulation.

In all cases, the output stage operates as a straight amplifier. For 27-29-Mc. output, a crystal in the 7-Mc. region is required, and the output circuit of the oscillator is tuned to the fourth harmonic.

Mobile Gear with Quick-Heating Filaments for 50 and 144 Mc.

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. 20-21 to 20-29 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. 20-21 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 20-22, 20-23, and 20-24, 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. 20-27.

Note the method of neutralization used in the final stage. The copper fin (designated as C_{16} in Fig. 20-26) visible in the rear view of the 144-Mc. unit is a device occasionally found necessary in tetrode amplifiers. In this

Typical Operating Conditions in the 50- and 144-Mc. Mobile Transmitters of Fig. 20-21 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	—

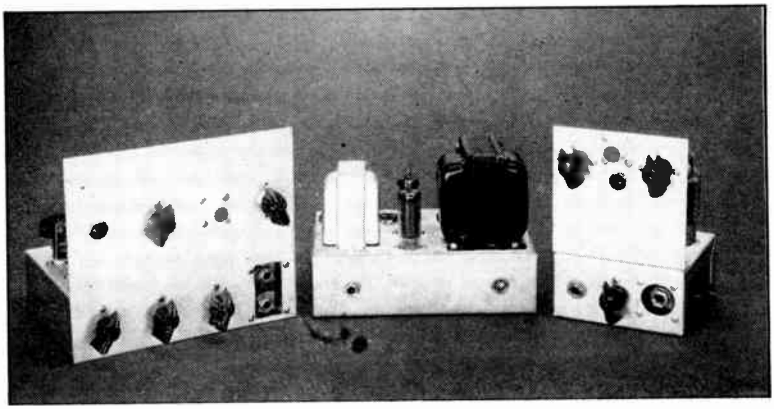

Fig. 20-21 — A complete mobile station for 50 and 144 Mc. using quick-heating filament tubes. The 144-Mc. r.f. section is at the left, the 50-Mc. portion at the right, and the modulator in the middle.

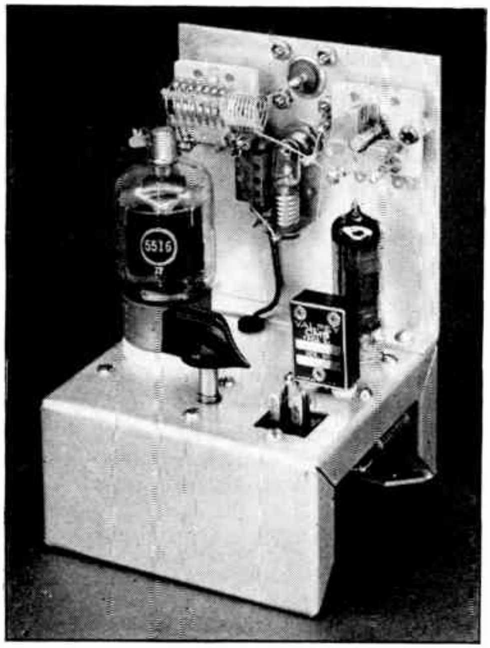

Fig. 20-22 — Rear view of the 50-Mc. r.f. section. The knob above the chassis is the cathode control. The final tank circuit is at the upper left, with antenna series tuning at the upper right.

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.

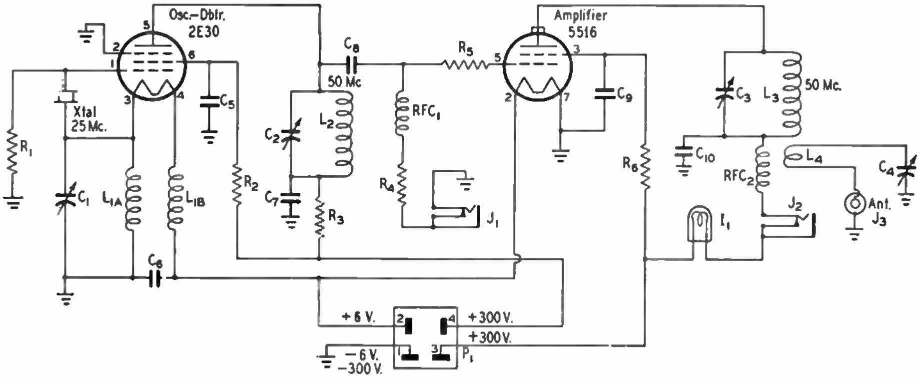

Fig. 20-23 — Schematic diagram of the 50-Mc. mobile unit.

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).

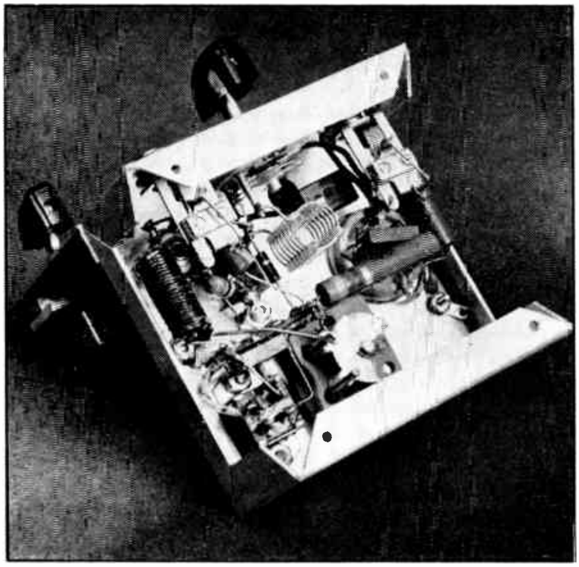

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Fig. 20-24 — Bottom view of the 50-Mc. rig. Note the interwound cathode coil at the left.

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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. 20-28 and 20-29, 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. 20-29, can be dispensed with.

The male power plug, P_1 in Fig. 20-29, 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

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Fig. 20-25 — Rear view of the 144-Mc. mobile unit. The copper fin at the side of the final tube is a neutralizing adjustment.

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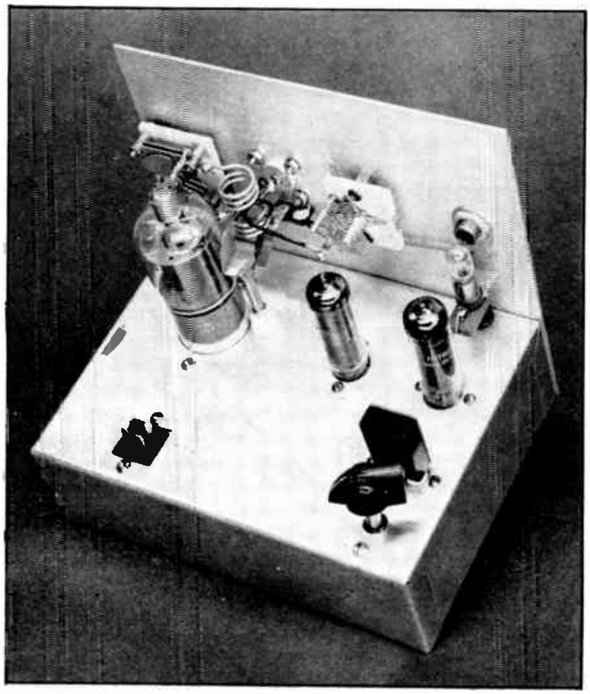

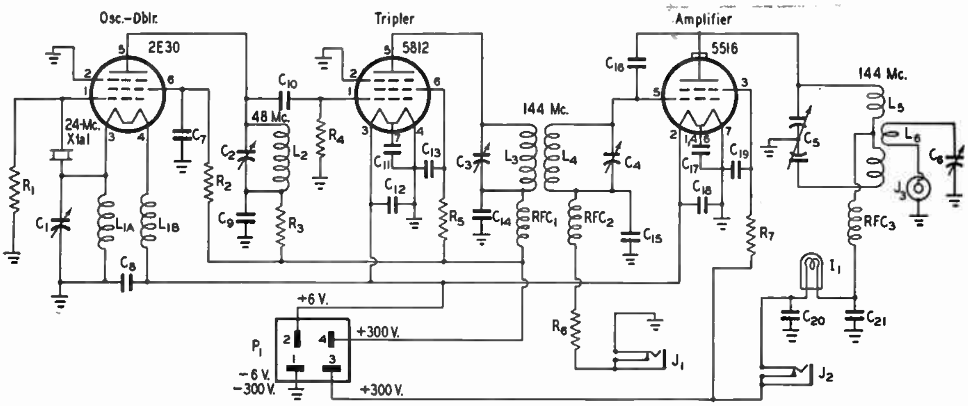

Fig. 20-26 — Schematic diagram of the 144-Mc. r.f. section.

C_1 — 50-μμfd. variable (Millen 20050).
C_2, C_3, C_4 — 15-μμfd. variable (Millen 20015).
C_5 — 6-μμfd.-per-section butterfly variable (Cardwell ER-6-BFS).
C_6 — 35-μμfd. 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-μμfd. mica.
C_{10} — 47-μμfd. mica.
C_{16} — Neutralizing-capacitor plate — see text and Fig. 20-25.
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. 20-27.)
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-

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Fig. 20-27 — 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.

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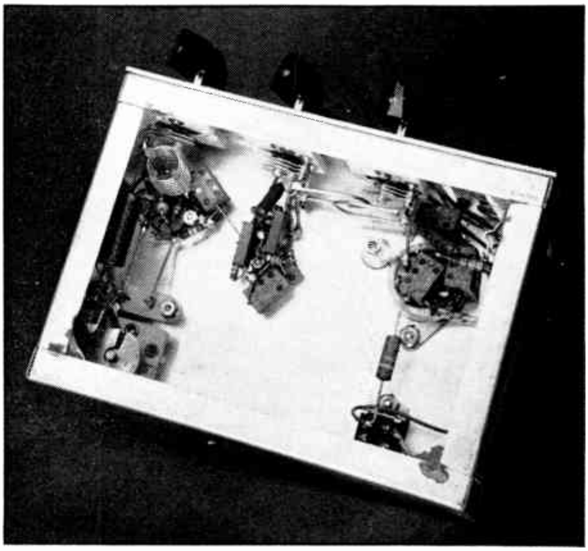

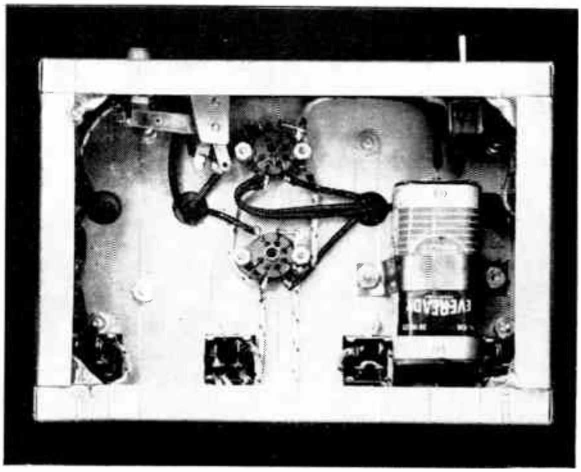

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Fig. 20-28 — Bottom view of the modulator and power-distribution unit.

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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.

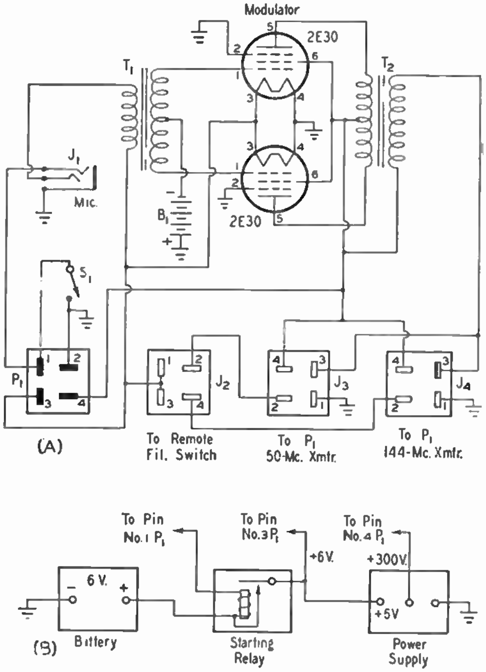


Fig. 20-29 — Schematic diagram of the modulator unit. Chassis size, 2 by 5 by 7 inches. Connections to the power plug and jacks on the unit are shown at A. External power circuits are given in B.

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).

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 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.

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.

Mobile Power Supply

By far the majority of amateur mobile installations depend upon the car storage battery as the source of power. The tube types used in equipment are chosen so that the filaments or heaters may be operated directly from the battery. High voltage may be obtained from a supply of the vibrator-transformer-rectifier type or from a small motor-generator operating from the battery.

Filaments

Because tubes with directly-heated cathodes (filament-type tubes) have the advantage that they can be turned off during receiving periods and thereby reduce the average load on the battery, they are preferred by some for transmitter applications. However, the choice of types with direct heating is limited, especially among those for 6-volt operation, and the saving may not always be as great as anticipated, because directly-heated tubes may require greater filament power than those of equivalent rating with indirectly-heated cathodes. In most cases, the power required for transmitter filaments will be quite small compared to the total power consumed.

Plate Power

Under steady running conditions, the vibrator-transformer-rectifier system and the motor-generator-type plate supply operate with approximately the same efficiency. However, for the same power, the motor-generator's overall efficiency may be somewhat lower because it draws a heavier starting current. On the other hand, the output of the generator requires less filtering and sometimes trouble is experienced in eliminating interference from the vibrator.

Mobile Power Considerations

Since the car storage battery is a low-voltage source, this means that the current drawn from the battery for even a moderate amount of power will be large. Therefore, it is important that the resistance of the 6-volt circuit be held to a minimum by the use of heavy conductors, no longer than necessary, and good solid connections. A heavy-duty relay should be used in the line between the battery and the plate-power unit. An ordinary toggle switch, located in any convenient position, may then be used for the power control. A second relay may sometimes be advisable for switching the filaments. If the power unit must be located at some distance from the battery (in the trunk, for instance) the 6-volt cable should be of the heavy military type.

A complete mobile installation may draw 30 to 40 amperes or more from the 6-volt battery. This requires a considerably increased demand from the car's battery-charging generator. The voltage-regulator systems on cars of recent years will take care of a moderate increase in demand if the car is driven fair distances regularly at a speed great enough to ensure maximum charging rate. However, if much of the driving is in urban areas at slow speed, or at night, it may be necessary to modify the charging system. Special communications-type generators, such as those used in police-car installations, are designed to charge at a high rate at slow engine speeds. The charging rate of the standard system can be increased within limits by tightening up on the voltage-regulator spring. This should be done with caution, however, checking for excessive generator temperature or abnormal sparking at the commutator. The average car generator has a rating of 35 amperes, but it may be possible to adjust the regulator so that the generator will at least hold even with the transmitter, receiver, lights, heater, etc., all operating at the same time.

Another scheme that has been used to increase generator output at slow driving speeds is to decrease slightly the diameter of the generator pulley. This means, of course, that the generator will be running above normal at high driving speeds. Some generators will not stand the higher speed without damage.

If higher transmitter power is used, it may be necessary to install an a.c. charging system. In this sytem, the generator delivers a.c. and works into a rectifier. A charging rate of 75 amperes is easily obtained. Commutator trouble often experienced with d.c. generators at high current is avoided, but the cost of such a system is rather high.

Some mobile operators prefer to use a separate battery for the radio equipment. Such a system can be arranged with a switch that cuts the auxiliary battery in parallel with the car battery for charging at times when the car battery is lightly loaded. The auxiliary battery can also be charged at home when not in use.

A tip: many mobile operators make a habit of carrying a pair of heavy cables five or six feet long, fitted with clips to make a connection to the battery of another car in case the operator's battery has been allowed to run too far down for starting.

Mobile Antennas

Most practicable mobile antenna systems are basically of the quarter-wave type, the car body serving as "ground." Exceptions are the half-wave systems sometimes used for 50 and 114 Mc. At 29 Mc., a simple quarter-wave vertical whip (approximately 8 feet) is feasible. A quarter-wave system for lower frequencies is simulated by the addition of loading inductance and capacitance to the 10-meter whip to make the system resonant at the operating frequency. Capacitive loading at the top end of the whip is preferable from the consideration of radiating efficiency, since it increases the current in the upper portion of the antenna more than any other loading system. This is desirable because that is the part of the antenna that is highest above ground and from which most-effective radiation takes place. However, top loading in most forms presents a mechanical problem, since it places a considerable weight at the top of the light whip structure. For this reason, it is often considered preferable to place the loading coil at the base of the whip antenna. More recently, a compromise has found favor in many mobile installations. In this arrangement, popularly called center loading, the loading coil is placed part way up on the antenna. The section of the whip above the loading coil serves as a loading capacitance as well as part of the radiating system.

A BASE-LOADED ANTENNA

The diagram of a base-loaded mobile antenna system is shown in Fig. 20-30. The capacitance of the antenna is combined with the inductance of the loading coil at the base to make the system resonant. The capacitance of a vertical antenna shorter than a quarter wavelength is given by:

$$C_a = \frac{17L}{\left[\left(log_e \frac{24L}{D}\right) - 1\right]\left[1 - \left(\frac{FL}{246}\right)^2\right]}$$

where

C_a = capacitance of antenna in $\mu\mu$fd.
L = antenna height in feet
D = diameter of radiator in inches
F = operating frequency in Mc.

$$log_e \frac{24L}{D} = 2.3\ log_{10} \frac{24L}{D}$$

The graph of Fig. 20-31 shows the capacitance of whips of various diameters and lengths. From this, the value of loading-coil inductance required to resonate at the desired frequency may be calculated.

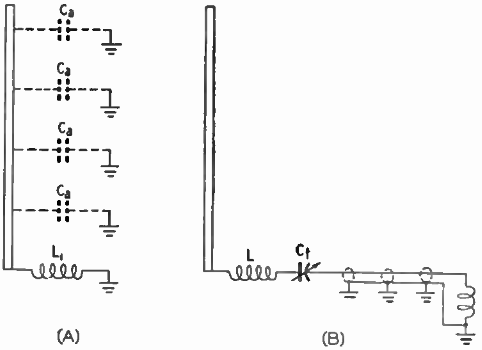


Fig. 20-30 — A — The capacitance of the vertical whip is combined with the inductance of the base coil to resonate at the desired operating frequency. B — The loading and coupling circuit. *L* is the loading coil. C_t is a 200-$\mu\mu$fd. variable condenser to tune out the reactance of the link pick-up coil.

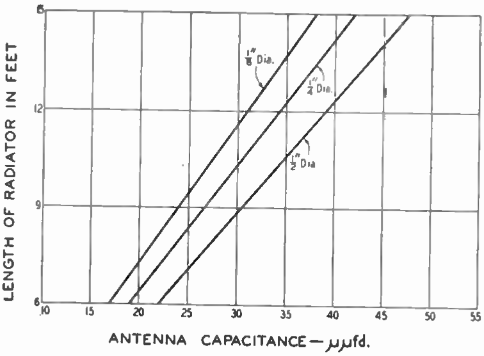


Fig. 20-31 — Graph showing the capacitance of short vertical antennas for various diameters and lengths.

Fig. 20-32 shows the installation of a base-loaded antenna on the rear apron of a car. The coil form is a wood dowel 1⅜ inches in diameter and 2 feet long. A hole 4 inches deep to fit the radiator snugly is drilled in the top end of the dowel. If the whip is hollow, it should be strengthened at the base by plugging it with dowel or aluminum rod for a distance of about 2 feet.

For 3.9 Mc., and a capacitance of 27 $\mu\mu$fd. (½-inch-diameter whip, 8 feet long), the form is close-wound with 100 turns of No. 14 enameled wire. The coil winding, which is about 8 inches long, is centered on the dowel form and the space at the bottom is used for mounting on the car. For 14 Mc., approximately 10 turns are required. A tap can be provided on the 75-meter coil so that all but 10 turns may be short-circuited.

As shown in Fig. 20-30, the antenna is fed with RG-8/U coaxial cable linked to the output tank coil. A variable condenser, C_t, is provided to tune out the reactance of the link. This condenser may be enclosed in a case and mounted near the base of the antenna, or placed at the other end of the line on the transmitter chassis.

Fig. 20-32 — The base loading coil and tuning condenser mounted on the rear apron of W2NNK's car.

● A TUNABLE CENTER-LOADED ANTENNA FOR 75 METERS

The installation of a center-loaded whip for 75 meters is shown in Figs. 20-33 and 20-34. To facilitate accurate adjustment of resonance, the coil is fitted with a movable brass slug that acts principally as a small variable capacitance, as shown in the sketch of Fig. 20-35. The coil shown has an inductance of 80 µh. It consists of 73 turns of No. 12 enameled wire close-wound on a 2-inch diameter bakelite tube. The tubing is boiled in household paraffin before winding, and the completed winding is dipped several times in the same substance to make it weather-resistant. Details of the assembly are shown in Fig. 20-35.

The coil is fastened to the top of a 27-inch length of ½-inch i.d. copper pipe. The ends of the pipe are plugged with brass inserts sweat-soldered in. The inserts are drilled and tapped for 5/16-18 machine screws. The portion of the antenna above the loading coil is a 40-inch collapsible whip with a 34-inch length of ⅛-inch brazing rod soldered to the top to make the overall length 74 inches. The entire assembly is designed to be fastened to the rear bumper apron with two ¼-inch bolts. On cars without this apron, a bumper fitting may be devised.

The antenna is fed with a length of RG-8/U coaxial cable. The coupling and change-over circuit is shown in Fig. 20-33. Keeping the output tank circuit tuned to resonance, the slug is adjusted until the plate current rises and passes through a peak. If this peak does not come up to rated loaded plate current, the coupling to the output tank should be increased.

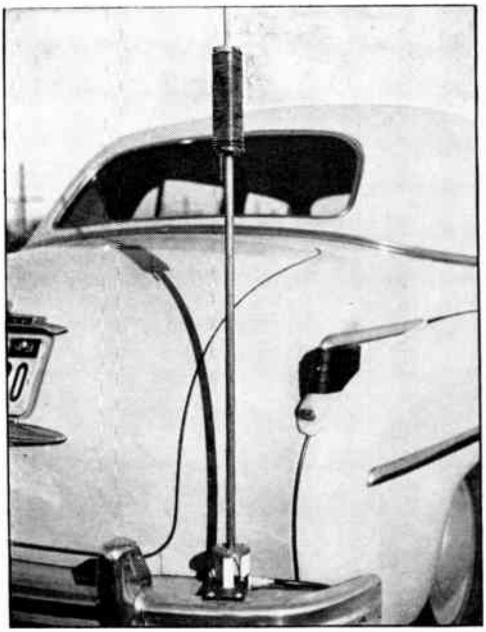


Fig. 20-34 — A close-up view of the bottom section of the center-loaded antenna of W2ABS, showing the mounting base and the loading coil. The tuning adjustment screw can be seen at the bottom end of the coil.

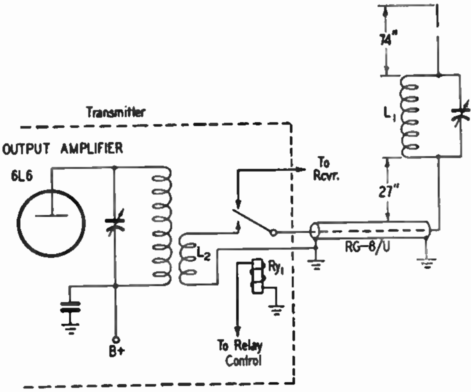


Fig. 20-33 — Electrical diagram of the slug-tuned mobile antenna.
C_1 — Tunable brass slug. See text and Fig. 20-34.
L_1 — 73 turns No. 12 enameled, close-wound on 2-inch diameter bakelite form.
L_2 — 4-turn coupling coil. See text.
Ry_1 — Antenna changeover relay.

● AN ALL-BAND ANTENNA

An antenna system that may be adjusted for operation in any band from 2 meters to 160 is illustrated in the sketch of Fig. 20-36. In the arrangement shown, a feed-through insulator is set in the roof of the car, near the rear. 144-50- and even 28-Mc. ¼-wave whips can be attached at this point and fed with a short coaxial line link-coupled directly to the output tank circuit. For the lower frequencies, a center-loaded system is used. As the sketch shows, the antenna for lower frequencies consists of a 12-inch and a 45-inch section of whip at the respective bottom and top ends of the loading coil. This assembly plugs into a 30-inch bakelite-tubing mounting. Also part of the radiating system is a 43-inch wire connecting the bottom of the whip to the feed-through insulator in the top of the car. This arrangement is fed from a parallel-tuned antenna tank circuit link-coupled to the output amplifier.

The bakelite mounting is 1¼ inches o.d. and ¾ inch inside. It is screwed on with standard iron-pipe fittings to a steel bracket bolted under the bumper guard. The top of the mounting is fitted with a Johnson No. 70 ⅜-inch jumbo banana jack so that the whip can be easily detached.

The loading coil consists of 200 turns of No. 22 d.s.c. wire, tapped every 5 turns, wound on a 12-inch section of 1¼-inch bakelite tubing. The turns are spaced the diameter of the wire and the taps are made by wrapping small soldering lugs around the wire and soldering them fast. The coil is then given two coats of lacquer to make it weather-resistant. Plugs made from 1-inch-diameter brass rod are fitted into both ends of the tubing. The plugs are drilled to

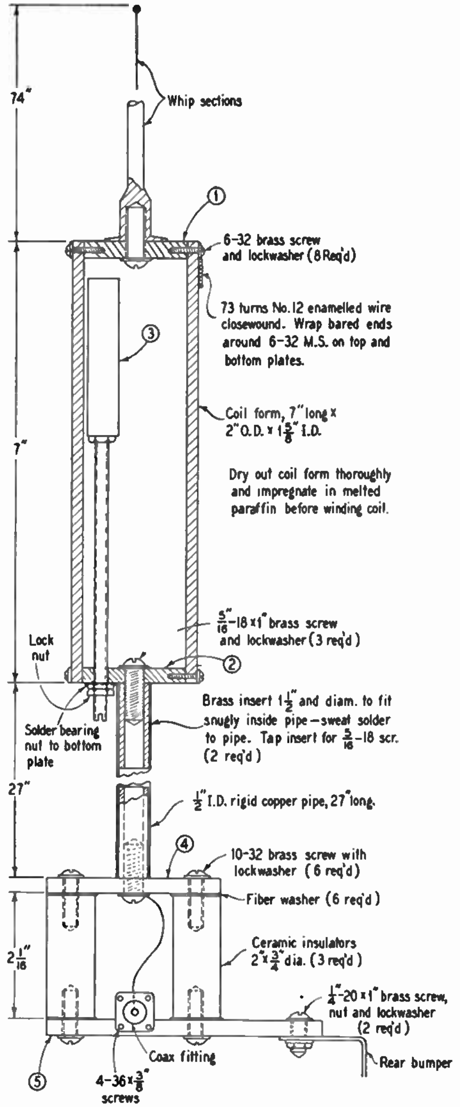


Fig. 20-35 — Detail sketch of the loading-coil form and mounting for the adjustable center-loaded mobile antenna. (1) and (2) are brass end plates. (3) is the tuning slug. (4) is the brass mounting plate. (5) is a mounting bracket made of aluminum.

take jumbo-size banana jacks. The antenna sections are of ⅜-inch rod to fit the jacks. The entire assembly may be taken apart in seconds.

For 10-meter operation of the bumper-mounted antenna, the loading coil is shorted out. The 20-meter band is covered with 5 to 10 turns, the remainder being shorted out. On 40, the system resonates with 35 to 60 turns, while 135 to 190 turns are required for the 75-meter band. The setting of the tap is critical for any given frequency. The system may be operated on 160 meters by making a second loading coil to plug into the top of the first. This should have 325 turns of No. 24 enameled wire, wound similarly to the first coil. No taps are necessary, however, adjustment being made on the first coil. The antenna resonates at 1820 kc. with the first coil set at 140 turns and at 1880 kc. with 80 turns.

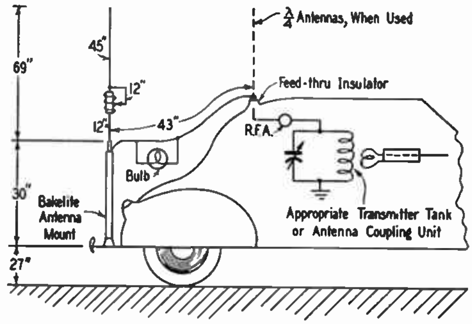
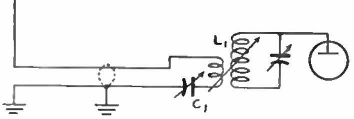


Fig. 20-36 — The all-band mobile antenna installation. As an alternative to feeding through an insulator in the roof of the car, the connection to the coupling circuit may be made through the bakelite tubing mount, when the transmitter is mounted in the car trunk.

ANTENNAS FOR 50 AND 144 MC.

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. 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. 20-37. 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.

Fig. 20-37 — Method of feeding quarter-wave mobile antennas with coaxial line. C_1 should have a maximum capacitance of 75 to 100 $\mu\mu$fd. for 28- and 50-Mc. work. L_1 is an adjustable link.



The short antenna required for 144 Mc. (approximately 19 inches) permits mounting the antenna on the top of the car. This provides good coverage in all directions, the car body acting as a ground plane. When the antenna is mounted elsewhere on the car, it is apt to show quite marked directivity. Because of this it is desirable to use the same antenna for both transmitting and receiving.

CHAPTER 21

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. The regulations also impose certain requirements with respect to plate-supply filtering, stability and purity of the transmitted signal, measurement of d.c. plate power input, 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.

● 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. 21-1 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.

WWV SCHEDULES

Standard radio and audio frequencies are broadcast continuously 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, 440 or 600
5.0	8.0	1, 440 or 600
10.0	9.0	1, 440 or 600
15.0	9.0	1, 440 or 600
20.0	8.5	1, 440 or 600
25.0	0.1	1, 440 or 600
30.0	0.1	1
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 440- and 600-cycle standard audio frequencies are transmitted in alternative five-minute periods, beginning with 600 c.p.s. in the first five-minute period of each hour.

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 (in GMT) is given in telegraphic code and in EST by voice. A station announcement is given in voice on the hour and half hour.

An announcement of radio propagation conditions is broadcast in code at 19 and 49 minutes past the hour. The letters transmitted have the following significance:

W — Ionospheric disturbance in progress or expected.
U — Unstable conditions expected.
N — No warning.

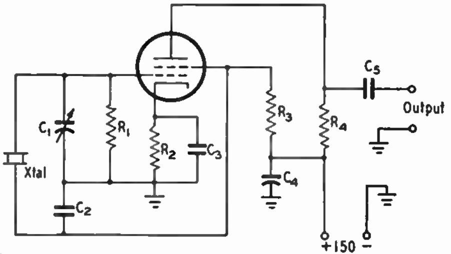


Fig. 21-1 — Circuit for crystal-controlled frequency standard. Tubes such as the 6SK7, 6SH7, 6AU6, etc., are suitable.

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.

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.

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 with 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. 21-1 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. 21-2. The mechanical construction should parallel that of the VFOs shown in the transmitting chapter. 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.

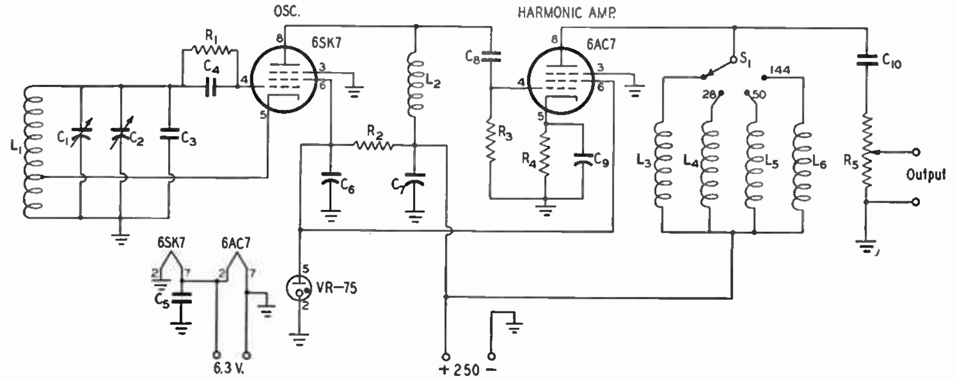


Fig. 21-2 — Heterodyne frequency meter with harmonic amplifier.

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. 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.

Fig. 21-3 — 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.

● 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. 21-3 to 21-6, 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. 21-4. 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

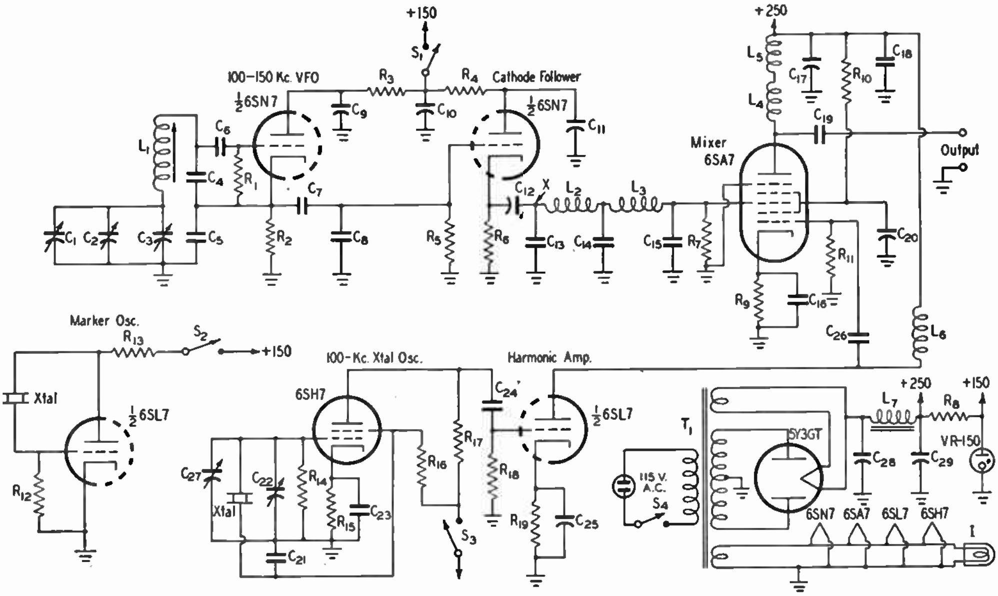


Fig. 21-4 — Circuit diagram of the additive frequency meter.

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.

Calibration

To set up the instrument, it is necessary first to adjust the VFO range exactly to 100–

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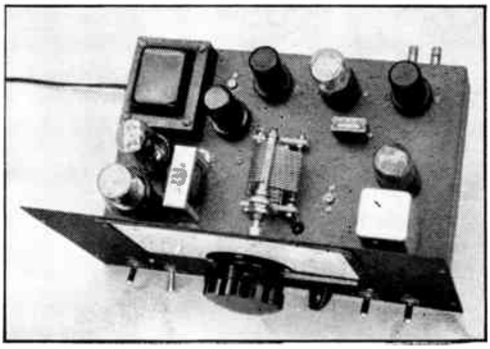

Fig. 21-5 — Chassis view of the additive frequency meter. Immediately in front of the power transformer are the rectifier and voltage-regulator tubes. The 100-kc. crystal, mounted in a metal-tube shell (James Knights), is just to the right of the power transformer. The tubes along the rear edge, from 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.

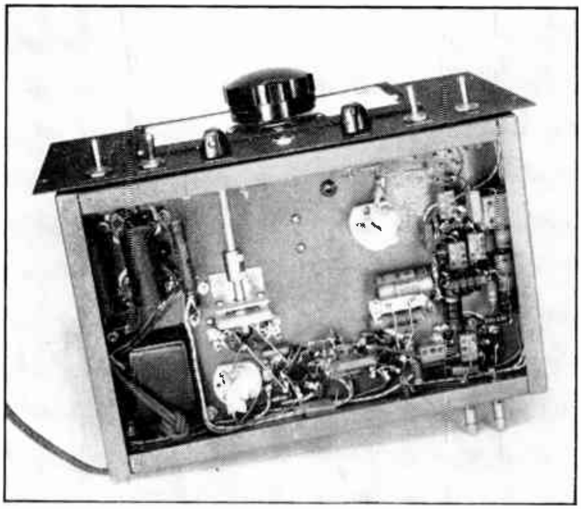

Fig. 21-6 — Bottom view of the frequency meter. Parts can be identified by reference to the tubes with which they are associated (see top view).

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. 21-7A. 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. 21-7B. 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

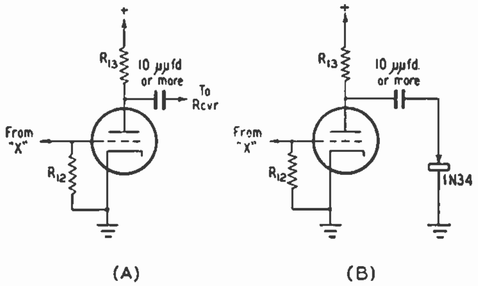

Fig. 21-7 — Temporary connections for amplifying VFO harmonics when calibrating. The marker-oscillator tube is used with the crystal removed.

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.

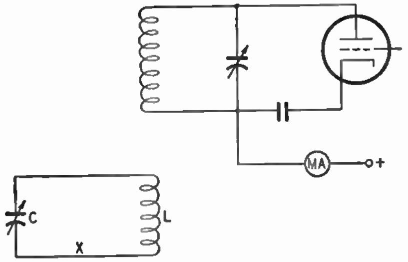

Fig. 21-8 — Absorption frequency meter and a typical application. The meter consists simply of the resonant circuit *LC*. When coupled to an amplifier or oscillator the tube plate current will rise when the frequency meter is tuned to resonance. The frequency may then be read from a calibrated condenser 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.

(A complete description of this system is given in May, 1949, *QST*.)

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-

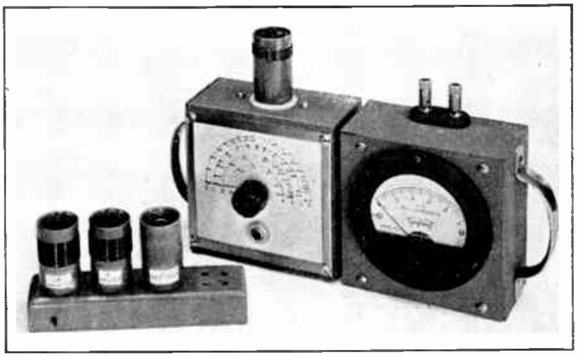

Fig. 21-9 — A sensitive absorption-type frequency meter with a crystal-detector rectifier and a d.c.-milliammeter indicating circuit. The meter is housed in a separate compartment so that it may be used with other measuring devices. The cabinet and front cover are drilled and tapped to accommodate the 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. 21-20.

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. 21-9 to 21-11, 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. 21-10, 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. 21-9 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.

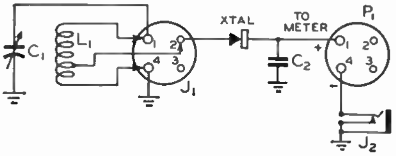

Fig. 21-10 — Circuit diagram of the absorption-type frequency meter.

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., ⅝ 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.

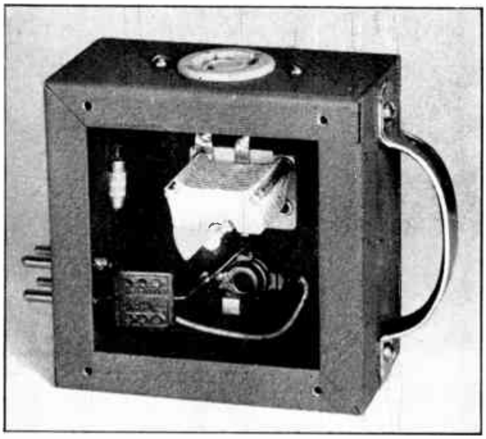

Fig. 21-11 — A rear view of the absorption-type frequency meter. The crystal is wired between the connector plug at the left and the coil socket at the top. The meter by-pass condenser is mounted between the plug and the grounded side of the 'phone jack. The variable-condenser terminals are connected directly to the coil socket.

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. 21-12). 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. 21-13. 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

$$F_{\text{Mc.}} = \frac{5905}{length\ \text{(inches)}}.$$

If the length is measured in meters,

$$F_{\text{Mc.}} = \frac{150}{length\ \text{(meters)}}.$$

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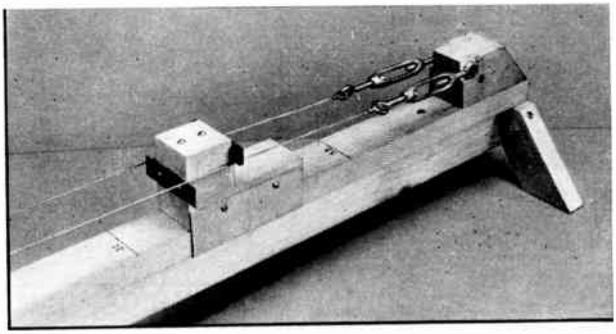

Fig. 21-12 — **One end of a typical Lecher wire system. The feet at each end keep the assembly from tipping over when in use. The wire is No. 16 bare solid-copper antenna wire (hard-drawn). The turnbuckles are held in place by a $\frac{3}{16}$ × 2-inch bolt through the anchor block. The other end of the line, the one connected to the pick-up loop, should be insulated.**

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.

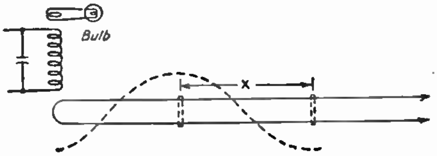

Fig. 21-13 — **Coupling a Lecher wire system to a transmitter tank coil. Typical standing-wave distribution is shown by the dashed line. The distance X between the positions of the shorting bar at the current loops equals one-half wavelength.**

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.

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. 21-14 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

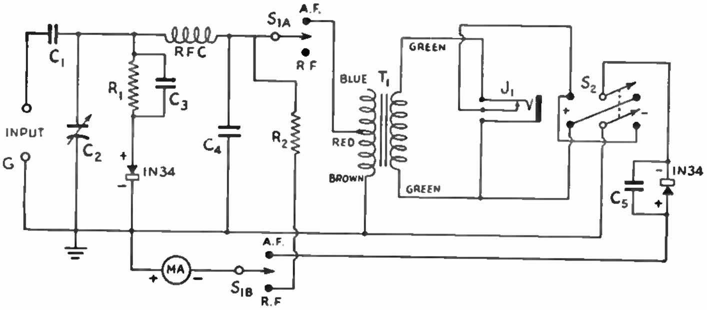

Fig. 21-14 — **Circuit of direct-reading modulation meter.**

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.

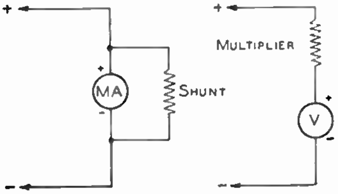

Fig. 21-15 — How voltmeter multipliers and milliammeter shunts are connected to extend the range of a d.c. meter.

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. 21-15 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 the data chapter 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.

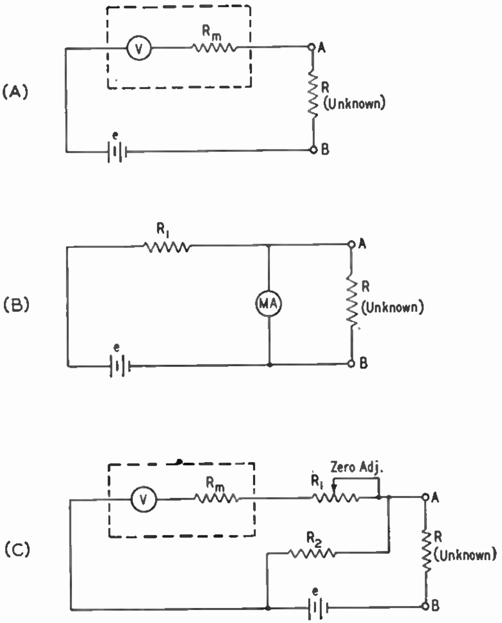

Fig. 21-16 — Circuits for measuring resistance. Values are discussed in the text.

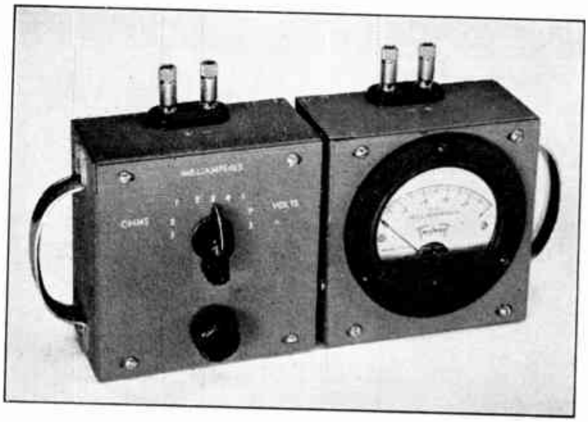

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Fig. 21-17 — An inexpensive multirange volt-ohm-milliammeter. The 2 × 4 × 4-inch cabinet at the left houses the multipliers, shunts, switch and zero-adjustment resistor. The meter is mounted in the metal cabinet shown at the right. The units are provided 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.

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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 many 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. 21-16A, 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. 21-16A is not suited to measuring low values of resistance (below a

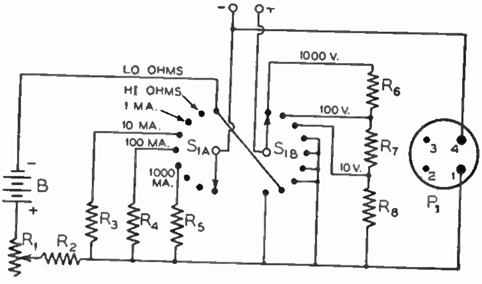

Fig. 21-18 — Diagram of the volt-ohm-milliammeter.

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).

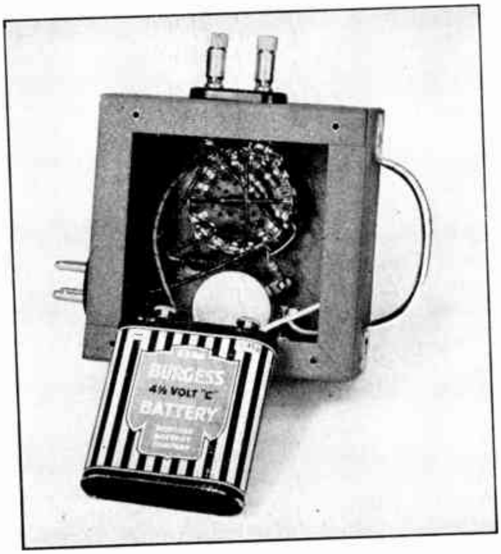

Fig. 21-19 — A rear view of the volt-ohm-milliammeter. The range-selector switch is mounted above the zero-adjustment potentiometer, and the shunts and multipliers are connected across the switch terminals. A four-prong male plug, for connection to the meter box, is shown at the left of the cabinet. 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. 21-16B 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_1 is at least 3000 ohms.

A third circuit for measuring resistance is shown in Fig. 21-16C. 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-ohms-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. 21-16C,

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 variable resistor is suitable with a 20,000-ohms-per-volt meter. The battery voltage is usually 3 volts for ranges up to 100,000 ohms or so and 6 volts for higher ranges.

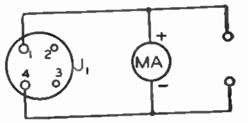

Fig. 21-20 — Wiring diagram of the 0–1 milliammeter shown in Figs. 21-9 and 21-17. J_1 is a 4-prong tube socket.

AN INEXPENSIVE V.O.M.

A combination multirange volt-ohm-milliammeter, reduced to simple and inexpensive terms, is shown in Figs. 21-17 to 21-20. Using a 0–1 milliammeter, the voltmeter has three

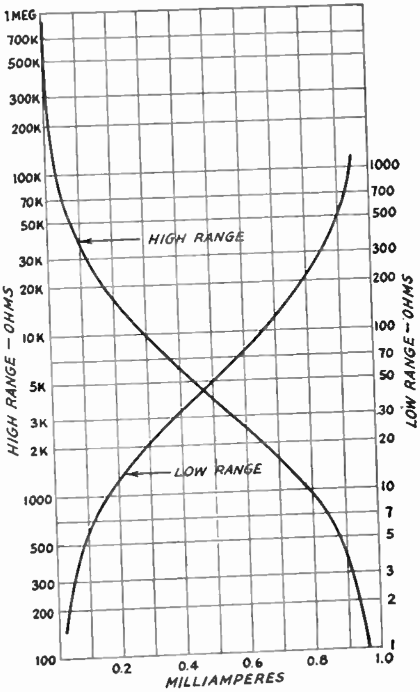

Fig. 21-21 — Calibration curve for the high- and low-resistance ranges of the volt-ohm-milliammeter.

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. 21-21. 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. 21-22 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. 21-23.

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. 21-22, 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

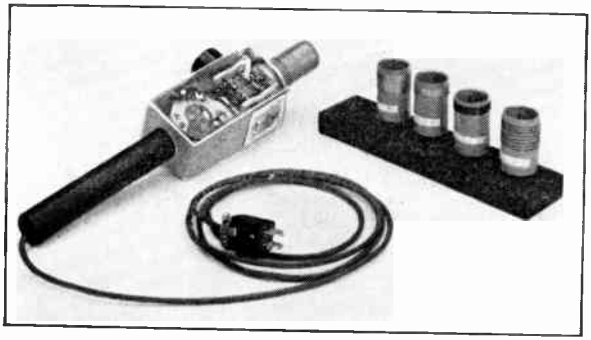

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Fig. 21-22 — Inexpensive grid-dip oscillator using a 955 and plug-in coils. The five coils shown cover the range 2850 kc. to 48 Mc. An external 0–1 d.c. milliammeter is used as an indicator. Power and meter connections are brought through the four-wire 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 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. 21-22, using a pointer on the rear shaft extension of the condensers as an indicator.

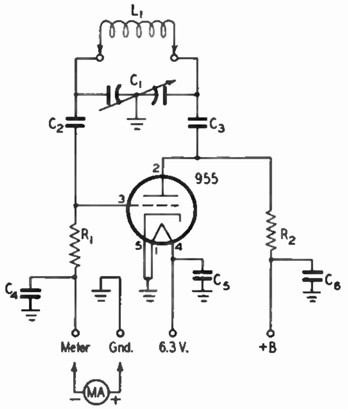

Fig. 21-23 — Circuit diagram of the grid-dip meter.

C_1 — Double-section midget, app. 42 μμfd. per section (Millen 21100 modified as described in text).
C_2, C_3 — 100-μμfd. 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.

◆

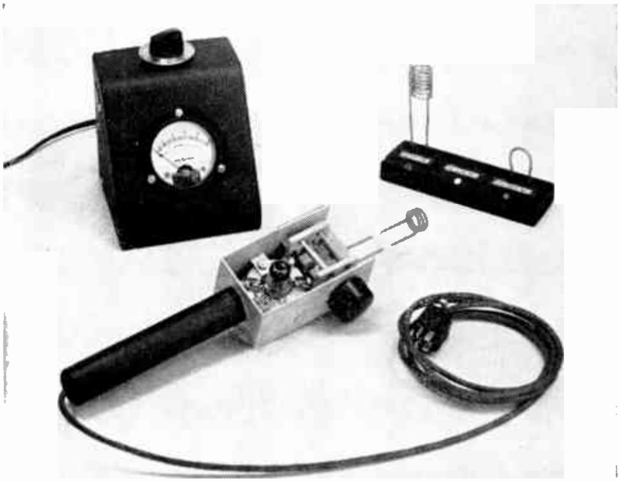

Fig. 21-24 — V.H.F. regenerative wavemeter/grid-dip meter, covering the 50–250 Mc. range. This is a high-sensitivity absorption-type wavemeter particularly useful for checking transmitter harmonics in television bands. The case in which the meter is mounted also contains the power supply. Regeneration 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. 21-27 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. 21-27, 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. 21-24 is similar in construction to the grid-dip meter of Fig. 21-22, 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

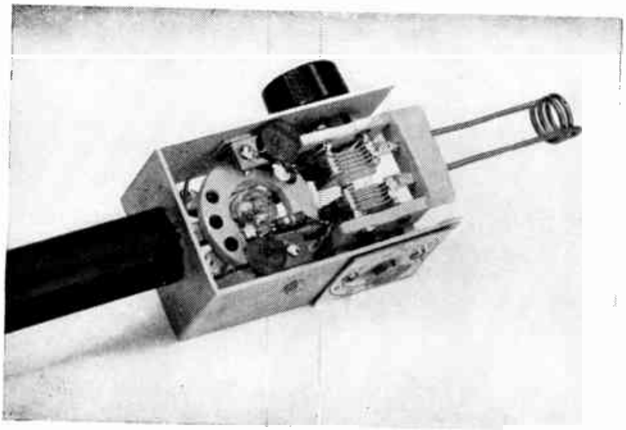

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Fig. 21-25 — A bottom view of the regenerative wavemeter/grid-dip meter. This view shows the bottom of the 955 socket, with the miniature tubular ceramics mounted between the stator sections of the tuning condenser and the grid and plate terminals 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. 21-26, 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. 21-27; 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. 21-27.

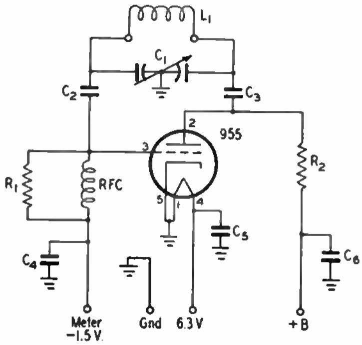

Fig. 21-26 — Circuit diagram of the regenerative wavemeter/grid-dip meter.

C_1 — Double-section midget, app. 36 μμfd. per section (Millen 21100 modified as described in text).
C_2, C_3 — 50-μμfd. ceramic (Centralab Hi-Kap).
C_4, C_5, C_6 — 0.001-μfd. ceramic (Sprague disc ceramic).
R_1 — 22,000 ohms, ½ watt, carbon.
R_2 — 68,000 ohms, ½ watt, carbon.
L_1 — 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.

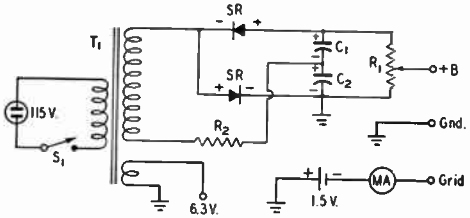

Fig. 21-27 — Power-supply circuit for the grid-dip meters shown in Figs. 21-22 and 21-24. When used with the meter of Fig. 21-22 the 1.5-volt battery should be omitted.

C_1, C_2 — 16-μfd. 150-volt electrolytic.
R_1 — 0.1-megohm potentiometer.
R_2 — 1000 ohms, 2 watts.
MA — 0–1 ma. (or smaller range for greater sensitivity).
S_1 — S.p.s.t. toggle (mounted on R_1).
SR — Selenium rectifier.
T_1 — 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. 21-22 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. 21-28. 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. 21-28. 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.

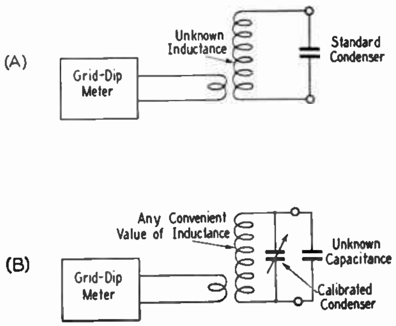

Fig. 21-28 — Set-ups for measuring inductance and capacitance with the grid-dip meter.

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.

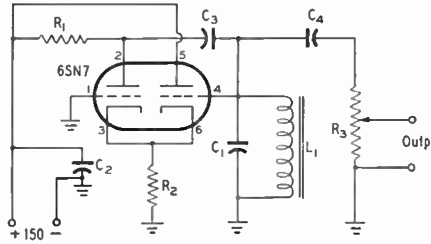

Fig. 21-29 — Audio oscillator circuit for fixed-frequency output.
C_1 — App. 0.05 μfd. (see text).
C_2 — 8-μfd. electrolytic.
C_3, C_4 — 0.1-μfd. paper.
R_1 — 68,000 ohms, 1 watt.
R_2 — 1500 ohms, 1 watt.
R_3 — 0.1-megohm potentiometer.
L_1 — App. 1 henry.

A circuit for a simple audio oscillator is given in Fig. 21-29. The second section of the 6SN7GT double triode is used to provide feedback 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-I.F. OSCILLATOR

For measurements requiring a variable-frequency audio source the signal generator shown in Figs. 21-30 to 21-33, inclusive, is relatively inexpensive and easy to build. It is also useful as an intermediate-frequency signal generator for aligning receiver i.f. circuits at any frequency up to 500 kc. The complete frequency range is 50 cycles to 500 kilocycles.

The oscillator consists of a 6AG7 amplifier coupled to a 6AG7 cathode follower. Two feedback loops are provided: (1) a cathode-to-cathode regenerative loop consisting of C_5 and lamp I_1; (2) a cathode-to-grid degenerative loop consisting of a bridged-T circuit. Oscillation occurs at the null frequency of the bridge, where the degeneration is minimum, and is determined principally by the values of C_6, C_7, C_8 and R_6 through R_{13}. The oscillator output is fed to the grid of a 6V6 cathode follower, which serves as an isolation stage between oscillator and load. Potentiometer R_{18} in the grid circuit controls the output voltage.

Output from the unit is taken across the 6V6 cathode resistor, R_{19}, through the coupling condenser, C_{11}. At 100 cycles the value given for C_{11} is suitable for working into load impedances as low as 20,000 ohms. For low audio frequencies and loads between 500 ohms and 20,000 ohms, excessive loss of voltage can be avoided by substituting a 25-μfd. electrolytic at C_{11}.

A 4-watt 115-volt lamp, I_1, regulates the feedback current and thus tends to keep the output voltage constant throughout the range. Potentiometer R_2 provides the means for adjusting the operating conditions to give minimum waveform distortion.

The 50-cycle to 500-kilocycle band is covered in four ranges, as follows:

Range	*Frequency*
A	50 to 500 kilocycles
B	5000 to 50,000 cycles
C	500 to 5000 cycles
D	50 to 500 cycles

Each step covers a 10-to-1 frequency range.

The ceramic trimmer, C_4, connected between the 6AG7 cathodes, has little effect at the lower frequencies, but to maintain the 10-to-1 frequency ratio on the high range this trimmer is essential.

The power supply uses a two-section choke input filter to insure good filtering. The com-

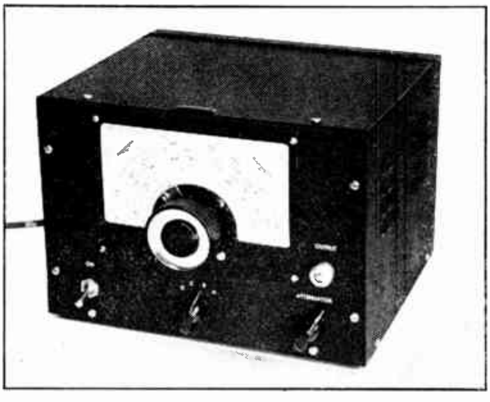

Fig. 21-30 — An *RC* oscillator covering the unusually wide range of 50 cycles to 500 kilocycles, with good waveform and practically constant output.

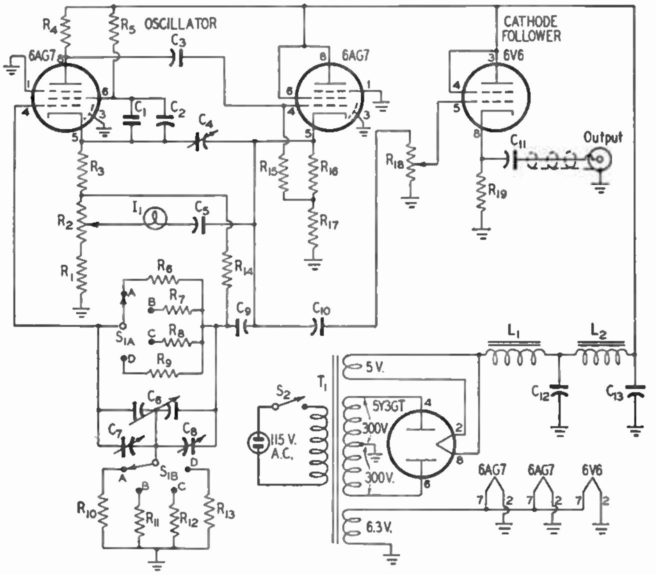

Fig. 21-31 — **Circuit diagram of the audio-i.f. test oscillator.**

C_1 — 0.002-μfd. mica.
C_2 — 40-μfd. 150-volt electrolytic.
C_3 — 1-μfd. 400-volt paper.
C_4, C_7, C_8 — 45-μμfd. ceramic trimmers (Centralab Type 822-BN).
C_5 — 100-μfd. 150-volt electrolytic.
C_6 — 500-μμfd.-per-section dual variable, broadcast receiver type.
C_9, C_{10}, C_{11} — 0.1-μfd. 400-volt paper.
C_{12}, C_{13} — 40-μfd. 450-volt electrolytic.
R_1 — 100 ohms, 1 watt.
R_2 — 2000-ohm wire-wound potentiometer.
R_3, R_{16} — 68 ohms, 1 watt.
R_4, R_{17} — 5000 ohms, 10 watts.
R_5 — 27,000 ohms, 2 watts.
R_6 — 15,000 ohms, ½ watt, 10%.
R_7 — 0.18 megohm, ½ watt, 10%.
R_8 — 1.8 megohms, ½ watt, 10%.
R_9 — 20.0 megohms, ½ watt, 10%.
R_{10} — 2700 ohms, ½ watt, 10%.
R_{11} — 39,000 ohms, ½ watt, 10%.
R_{12} — 0.33 megohm, ½ watt, 10%.
R_{13} — 3.3 megohms, ½ watt, 10%.
R_{14}, R_{15} — 1.0 megohm, 1 watt.
R_{18} — 0.5-megohm potentiometer.
R_{19} — 2200 ohms, 1 watt.
L_1, L_2 — 10-hy. 50-ma. chokes.
I_1 — 1-watt 115-volt lamp.
J_1 — Shorting-type microphone jack (Amphenol 73-CL PC1M).
S_1 — Single-section 2-pole 4-position ceramic.
S_2 — S.p.s.t. toggle switch.
T_1 — 300-0-300 v., 50 ma.; 5 v., 2 amp.; 6.3 v., 3 amp.

ponents are confined to the extreme rear of the chassis and shielded wire is used for the filament wiring.

Construction

The complete unit is housed in a standard 8 × 10 × 8-inch steel cabinet. The chassis is 7 × 9 × 2 inches.

The power transformer, T_1, is submounted at the rear of the chassis. The can-type electrolytics, C_{12} and C_{13}, are mounted above the chassis while the filter chokes are placed below.

The main tuning condenser, C_6, must be insulated from the chassis. Small porcelain stand-offs or a slab of polystyrene or bakelite sheet will be satisfactory. An insulated coupling must be used between the condenser and dial. The frequency-determining resistors, R_6 through R_{13}, are mounted on the ceramic range switch, S_1, which is located under the tuning control. These resistors must have the designated values or the frequency ranges will differ from those given. Resistors of 10 per cent tolerance are satisfactory.

On the front panel there are four controls and the output terminal. A National type SCN dial is used for tuning. In the lower corner of the panel is a toggle switch, S_2, for the a.c. line. The band-changing switch is placed under the tuning knob. At the lower right is the attenuation control, R_{18}. Just above this control is the output connector, J_1. These controls fasten the panel to the chassis.

Preliminary Adjustment

An oscilloscope should be used for adjusting the waveform and for calibrating the low-frequency ranges. Connect the output of the oscillator to the vertical plates of the 'scope and, with the range selector in position *D* and the tuning condenser, C_6, nearly at maximum, adjust the internal horizontal sweep in the 'scope for synchronization.

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Fig. 21-32 — **In this rear view of the oscillator the metal tube on the left is the first cathode follower. The tuning condenser and its trimmers are mounted on a piece of bakelite to insulate them from ground and the condenser is driven through an insulated coupling. The control shaft of the waveform potentiometer, R_2, is visible on the chassis to the right of the tuning condenser.**

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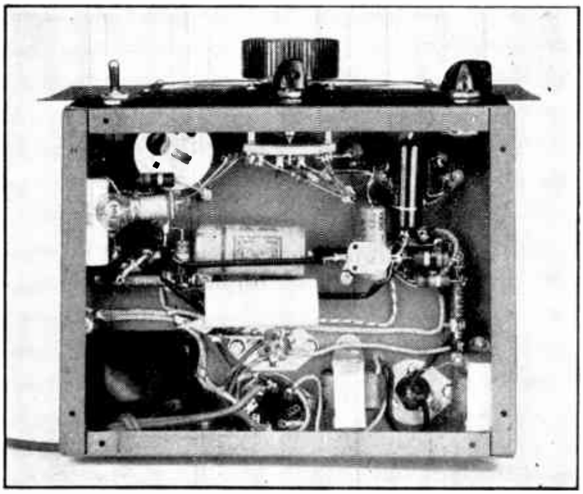

Fig. 21-33 — Bottom view of the audio-i.f. test oscillator. The filter chokes are at the bottom right. The frequency-determining resistors are supported by the ceramic range switch at the top center.

R_2 should be adjusted to give a good sine wave. In case the 'scope has no internal sweep, an external source of 60 cycles from a filament transformer can be used as the horizontal sweep, and the tuning condenser of the test oscillator adjusted until a single-loop Lissajous pattern appears. The pattern will resemble either a circle, ellipse, or straight line. Adjustment of R_2 will affect the symmetry of the loop about its own axes and the distortion will be least when the loop is perfectly symmetrical.

To adjust the ranges, set the tuning condenser approximately 10 dial divisions from minimum capacity with S_1 on range *D*. Trimmers C_7 and C_8 should be set to full capacity. Connect the output of the oscillator to the vertical plates of the 'scope. Feed the audio output of a receiver tuned to WWV to the horizontal plates. WWV sends either a 440- or 600-cycle tone, so make sure that the adjustment is made during the 440-cycle period. Adjust trimmers C_7 and C_8 a little at a time, keeping their capacities about equal, until a single-loop Lissajous figure is seen on the screen. This adjustment sets the high end of range *D* and at the same time fixes ranges *B* and *C*.

Most 'scopes are useless for calibration in the r.f. range. A simple yet effective method for adjusting the high end of range *A* utilizes a receiver calibrated over the broadcast band. For preliminary adjustments, the 500-kc. intervals starting at 1 Mc. are needed. However, the 10-kc. points from 600 kc. and up will be useful later on for calibration. Broadcast stations can be used to spot frequencies on the dial. By interpolation, the 10-kc. points can be marked with reasonable accuracy. A 10-kc. multivibrator would be excellent for calibration, but the station spotting method will give very satisfactory results. After calibrating the receiver, the output of the oscillator should be connected to the antenna terminals through a shielded cable. Set R_{18} at maximum and the main tuning dial five divisions from minimum capacity. With the receiver set at exactly 1000 kc. and the b.f.o. in the "on" position, adjust trimmer C_4 for zero beat. The oscillator will be on 500 kc. if beats are observed *only* at 1000 kc. and 1500 kc. It may be necessary to try a few settings of C_4 before the right one is found.

Calibration

Up to 5000 cycles, covered by ranges *C* and *D*, the oscilloscope and the WWV standard audio signal are used for calibrating. Information on using Lissajous figures is given later in this chapter. Assuming that 60 cycles from the power line and WWV's 440- and 600-cycle tones are the standard signals available, it is feasible to calibrate up to 6000 cycles; above this frequency the patterns are too complex for rapid analysis.

Between 6000 and 10,000 cycles, the most feasible method is to obtain the points from a regular calibrated audio oscillator. Alternatively, a fixed-frequency oscillator (such as the simple type described earlier in this section) can be constructed in temporary fashion and adjusted to, say, 2000 cycles and used for obtaining points at 2-kc. intervals between 6 and 10 kc. by the Lissajous-figure method.

To spot points from 10 kc. to 500 kc., the full output of the oscillator on range *C* is fed into the calibrated receiver antenna terminals, and the tuning control should be adjusted until the signals fall at every 10-kc. point through the broadcast band. At this setting the oscillator frequency will be 10 kc. Considerable care, and several attempts, will undoubtedly be necessary before the correct setting is reached. The harmonic method described earlier in this chapter in the section on frequency measurement can be used for calibrating up to 500 kc.

In using the instrument, a warm-up period of about 20 minutes should be allowed for the frequency to stabilize. At the setting of R_2 that gives good waveform, the output with R_{18} at maximum is approximately 10 volts r.m.s. The attenuator gives smooth output control and is readily adjustable to outputs in the microvolt region even at 500 kc.

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. 21-34, 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 electric 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. 21-34. 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. 21-35 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

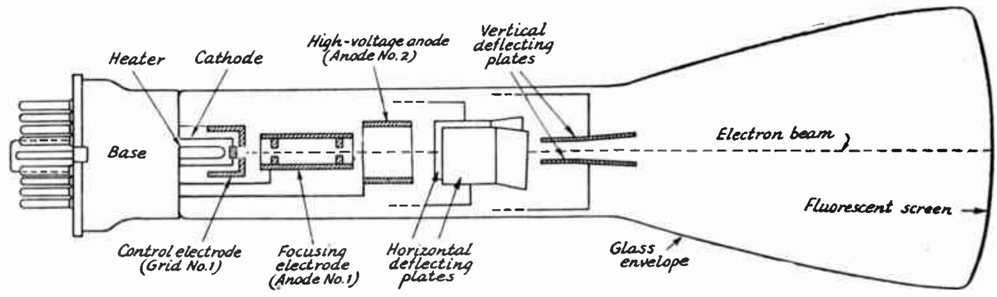

Fig. 21-34 — Typical construction for a cathode-ray tube of the electrostatic-deflection type.

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. 21-35, 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 the desired trace AH, at least at most frequencies within the audio range. The fly-back time is somewhat exaggerated in Fig. 21-35, 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.

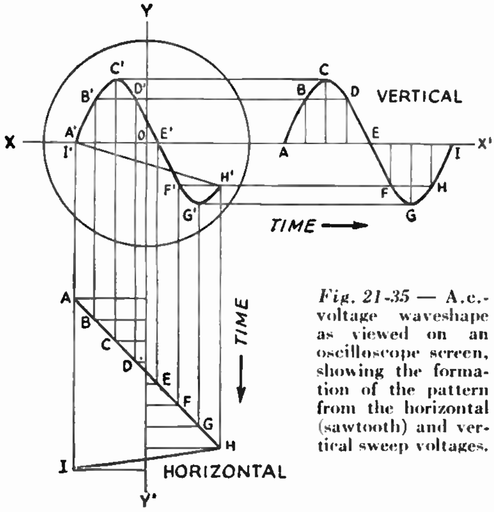

Fig. 21-35 — A.c.-voltage waveshape as viewed on an oscilloscope screen, showing the formation of the pattern from the horizontal (sawtooth) and vertical sweep voltages.

The 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 the chapter on amplitude modulation) 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. 21-36. Patterns of the type shown in Fig. 21-36 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 vertical deflecting plates. Should the lower frequency be applied to the vertical 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 1 to 3. 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 3 to 4. Assuming that the known frequency is applied to the horizontal plates, the unknown frequency is

$$f_2 = \frac{n_2}{n_1} f_1$$

where f_1 = known frequency applied to horizontal plates,

f_2 = unknown frequency applied to vertical plates,

n_1 = number of loops along a vertical edge, and

n_2 = number of loops along a horizontal edge.

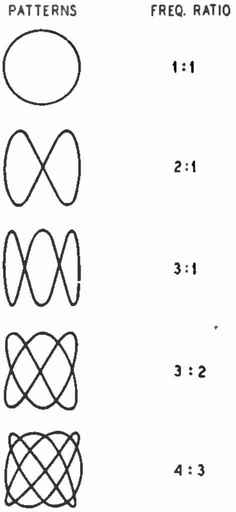

Fig. 21-36 — **Lissajous figures and corresponding frequency ratios for a 90-degree phase relationship between the voltages applied to the two sets of deflecting plates.**

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- and 600-cycle modulation on the WWV transmissions. 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 600 cycle voltages from the WWV signal can be taken from the headphone jack on a receiver. It is possible to calibrate over a 10-to-1 range, both upwards and downwards, from each of the latter frequencies and thus cover the audio range useful for voice communication.

● A SIMPLE OSCILLOSCOPE FOR MODULATION MONITORING

Figs. 21-37 through 21-39 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

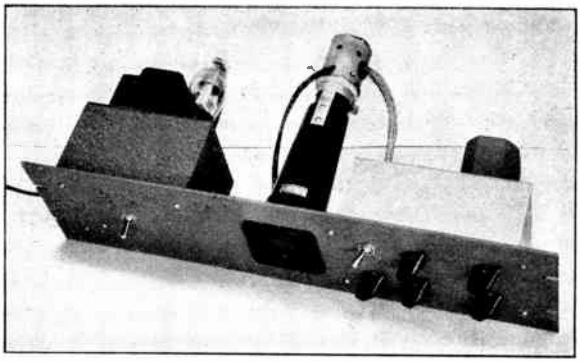

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Fig. 21-37 — Front view of a rack-mounting oscilloscope for modulation monitoring. All components are mounted on the rear of a 19 × 5½-inch rack panel. The power-supply components are built into a utility box bolted on the left side of the panel, and 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.

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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 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 the chapter on amplitude modulation. An external resistor, R_E in Fig. 21-38, must be used in series with the lead to the horizontal input

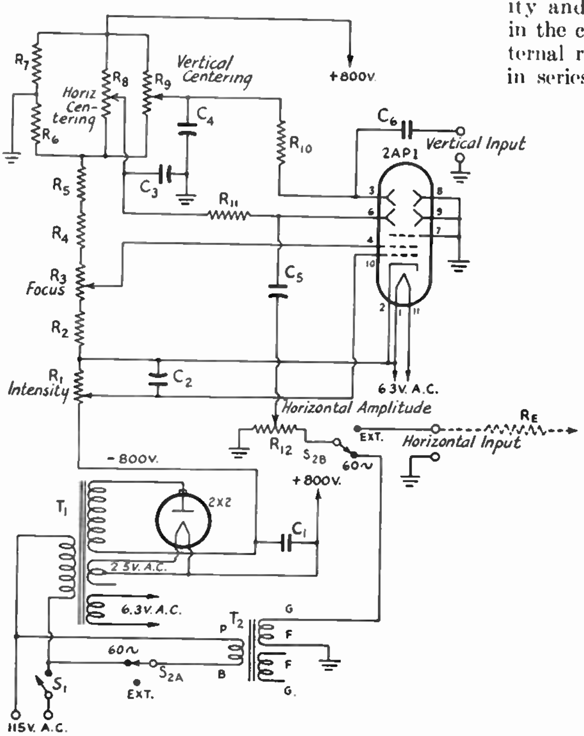

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Fig. 21-38 — Circuit diagram of the simple oscilloscope for modulation monitoring.

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.)

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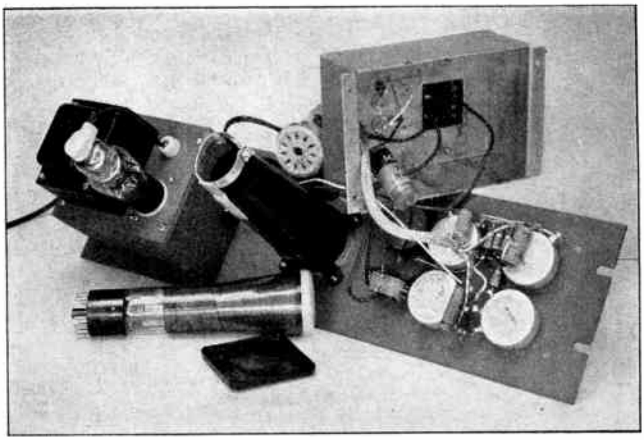

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Fig. 21-39 — Rear view of the rack-mounting oscilloscope. The shield covering the voltage-divider components has been removed to show wiring. Mounted on the shield are the audio transformer and the horizontal input terminals. The 'scope tube and its socket have been removed.

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terminals to reduce the audio voltage to the desired level, as described in the amplitude-modulation chapter.

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. 21-40. 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. 21-41, 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

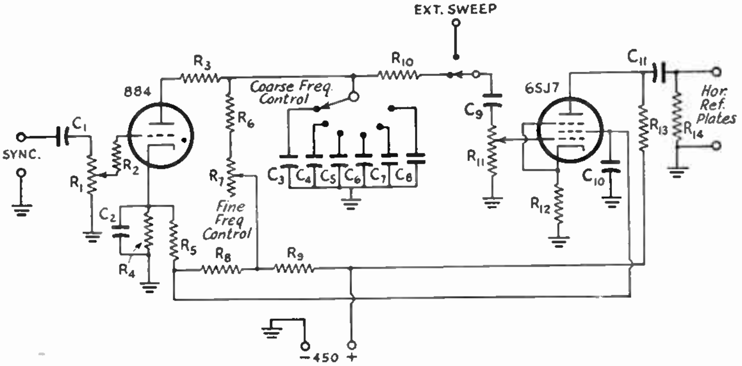


Fig. 21-40 — Linear sweep generator and horizontal amplifier.

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 — 0.04-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, 450 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.

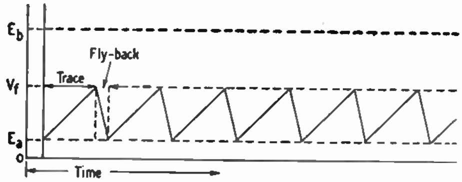


Fig. 21-41 — Condenser charging curves showing how a sawtooth wave is produced by a gaseous-tube linear sweep oscillator.

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. 21-40 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. 21-38 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. 21-42. It will be recognized as being practically similar to the "horizontal" amplifier of Fig. 21-40. 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. 21-38, 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. 21-38. 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.

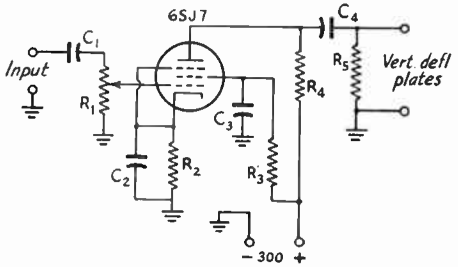


Fig. 21-42 — Circuit diagram of a vertical amplifier for an oscilloscope.

C_1, C_3, C_4 — 0.1-μfd. paper, 100 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 and Transmission-Line Measurements

Two principal types of measurements are made on antenna systems: 1) The standing-wave ratio on the transmission line, as a means for determining whether or not the antenna is properly matched to the line; 2) The comparative radiation field strength in the vicinity of the antenna, as a means for checking the directivity of a beam antenna and as an aid in adjustment of element tuning and phasing. Both types of measurements can be made with rather simple equipment.

● FIELD STRENGTH MEASUREMENTS

The radiation field of an antenna is measured with a device that is essentially a very simple receiver equipped with an indicator to give a visual representation of the comparative signal strength. Such a **field-strength meter** is used with a "pick-up antenna", which should always have the same polarization as the antenna being checked — e.g., the pick-up antenna should be horizontal if the transmitting antenna is horizontal. Care should be taken to prevent stray pick-up by the field-strength meter itself or by any transmission line that may connect it to the pick-up antenna.

Field-strength measurements preferably should be made at a distance of several wavelengths from the transmitting antenna being tested. Measurements made within a wavelength of the antenna may be misleading, because of the possibility that the measuring equipment may be responding to the combined induction and radiation fields of the antenna, rather than to the radiation field alone. Also, if the pick-up antenna has dimensions comparable with those of the antenna under test it is likely that the coupling between the two antennas will be great enough to cause the pick-up antenna to tend to become part of the radiating system and thus result in misleading field-strength readings.

A desirable form of pick-up antenna is a dipole installed at the same height as the antenna being tested, with low-impedance line such as 75-ohm Twin-Lead connected at the center to transfer the r.f. signal to the field-strength meter. The length of the dipole need only be great enough to give adequate meter readings. A half-wave dipole will give maximum sensitivity, but such length will not be needed unless the distance is several wavelengths and a relatively insensitive meter is used.

Field-Strength Meters

The crystal-detector wavemeter described earlier in this chapter may be used as a field-strength meter. It may be coupled to the transmission line to the pick-up antenna by means of a link of a few turns wound around the wavemeter coil. Also, the wavemeter proper may be connected to the milliammeter through a section of lampcord or similar two-conductor cable of any convenient length. This permits the milliammeter unit to be near the point where adjustments are being made, even though the pick-up antenna and wavemeter may be several wavelengths away.

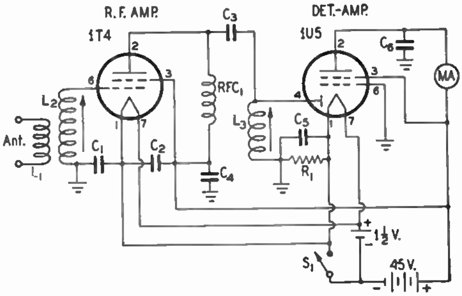


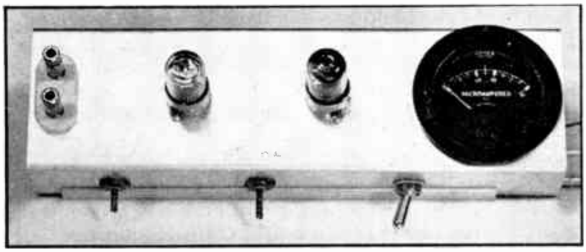
Fig. 21-44 — **Wiring diagram of the sensitive field-strength meter.**

C_1, C_2, C_6 — 0.001-μfd. ceramic.
C_3, C_5 — 470-μμfd. ceramic.
C_4 — 0.005-μfd. ceramic.
R_1 — 1.5 megohms.
L_1 — 14 Mc.: 8 turns No. 30 d.c.c.
28 Mc.: 6 turns No. 22 d.c.c.
L_2 — 14 Mc.: 34 turns No. 30 d.c.c.
28 Mc.: 24 turns No. 22 d.c.c.
L_3 — 14 Mc.: 27 turns No. 28 d.c.c.
28 Mc.: 16 turns No. 20 d.c.c.

L_1 wound over ground end of L_2. L_2 and L_3 close-wound on National XR-50 slug-tuned coil forms.

RFC_1 — 750 μh. (National R33).
S_1 — S.p.s.t. toggle.
MA — 0.5 milliammeter.

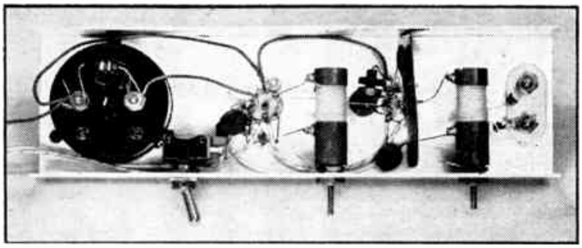
The indications with a crystal wavemeter connected as shown in Fig. 21-10 will tend to be "square law" — that is, the meter reading will be proportional to the square of the r.f. voltage. This exaggerates the effect of relatively small adjustments to the antenna system and gives a false impression of the improvement secured. The meter reading can be made more linear by connecting a fairly large resistance in series with the milliammeter (or microammeter). About



◆

Fig. 21-43 — **A logarithmic field-strength meter of high sensitivity. It uses two miniature battery-operated tubes and a 0-500 microammeter, and gives readings that are approximately proportional to the change in field strength in decibels.**

◆



◆

Fig. 21-45 — The logarithmic f.s. meter is constructed on a small aluminum channel. A small copper plate between the two coils is used for reducing the interstage coupling to the point where the r.f. amplifier is nonregenerative.

◆

10,000 ohms is required for good linearity. This considerably reduces the sensitivity of the meter, but the lower sensitivity can be compensated for by making the pick-up antenna sufficiently large.

A Sensitive Logarithmic F.S. Meter

For indicating the effect of antenna adjustments at a distant station, a logarithmic type of indicator is desirable in the field-strength meter since the meter readings with such an instrument are directly proportional to decibels. Figs. 21-43 to 21-45, inclusive, show a meter of this type. It makes use of the fact that the rectified d.c. output of a detector following a.v.c.-controlled r.f. stages tends to be logarithmic with respect to the r.f. voltage applied to the receiver.

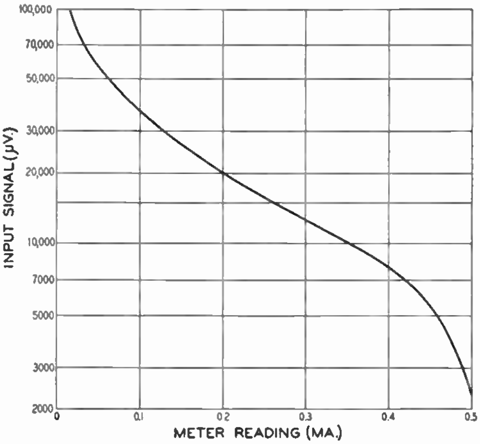


Fig. 21-46 — Typical calibration curve of the logarithmic field-strength meter. The curve is sufficiently logarithmic, for practical purposes, between about 0.05 and 0.45 ma. The way in which the readings vary with applied signal, and not the absolute value of the signal, is the important point, and since this will not change significantly so long as the same circuit is used, the curve above may be used with any similar instrument.

As shown in Fig. 21-44, the circuit includes an r.f. amplifier, a detector, and a d.c. amplifier, using miniature battery tubes. The rectified r.f. voltage developed across R_1 in the diode circuit of the 1U5 is applied through the ground connection to the grid of the 1T4 r.f. amplifier and thus controls its gain. The 1½-volt "A" battery is not connected to ground but is allowed to "float", permitting the a.v.c. voltage to be effective on the grids.

In the unit shown in the photographs, slug-tuned coils are used because of their small size and because they eliminate the need for variable tuning condensers. However, ordinary condenser-tuned circuits can be substituted; the only requirement is that the circuits must be tunable to the frequency at which the antenna is being adjusted. The only critical point about the construction of such a meter is to lay out the tuned circuits so that the r.f. amplifier is stable; otherwise, any convenient layout may be used.

With the values shown in Fig. 21-44 the no-signal plate current should be very close to 0.5 milliampere. A less-sensitive d.c. instrument will require more "B" voltage. Whatever the type of meter, the current may be brought to exactly full scale, with no signal input, by shunting it with a variable resistor of suitable range, depending on the internal resistance.

Fig. 21-46 is a typical calibration curve. The readings are approximately logarithmic over about 70 percent of the scale, with a range of about 20 db. Used with a folded-dipole pick-up antenna, the instrument is sensitive enough for use a few thousand feet away from a beam antenna fed with a few hundred watts.

● 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. Also, 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.

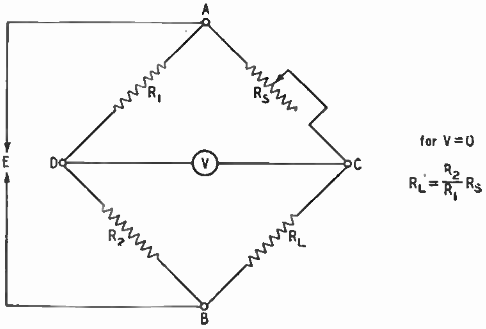


Fig. 21-47 — **Resistance bridge as used for resistance measurement. This fundamental circuit is the basis for one type of bridge for measuring standing-wave ratio.**

An alternative method uses a bridge circuit to measure the standing-wave ratio. While there are many forms of bridge circuits, the simple resistance bridge shown in Fig. 21-47 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,

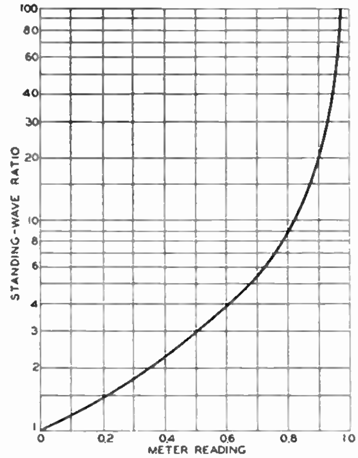


Fig. 21-48 — **Standing-wave ratio in terms of meter reading (relative to full scale) after setting outgoing voltage to full scale. This graph is a plot of the formula**

$$S.W.R. = \frac{V_o + V_r}{V_o - V_r}$$

where V_o and V_r are the outgoing and reflected components, respectively, of the voltage on the transmission line.

$$R_L = R_S \frac{R_2}{R_1}.$$

R_1 and R_2 are called the "ratio arms" of the bridge.

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 a matched 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 voltage reflected back from the far end of the line will appear at the terminals of the bridge and will register on the voltmeter. The relationship between voltmeter reading (in percentage of full scale) and standing-wave ratio is shown in Fig. 21-48. This curve applies *only* when the voltmeter impedance is extremely high — 20 times or more — compared with the impedance for which the bridge is designed.

While other bridge circuits can be used for s.w.r. measurement, the resistance bridge is about the simplest and easiest to build. It lends itself well to construction for coaxial lines and when so designed can be used for measurement of open-wire lines as shown later in this chapter.

● S.W.R. INDICATOR FOR COAXIAL LINES

Figs. 21-49 to 21-51, inclusive, illustrate the type of construction that should be used in a coaxial-line s.w.r. indicator. Coupling between various parts of the r.f. circuits should be as small as possible. Short leads in the r.f. wiring are 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. The loading resistor, R_1, places a constant low-resistance load on the transmitter and thereby helps maintain constant voltage across the bridge regardless of the load that may be connected to the output terminals. A 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 meter. This line voltmeter is a convenience in making measurements, because it will show whether or not the line voltage varies when shift-

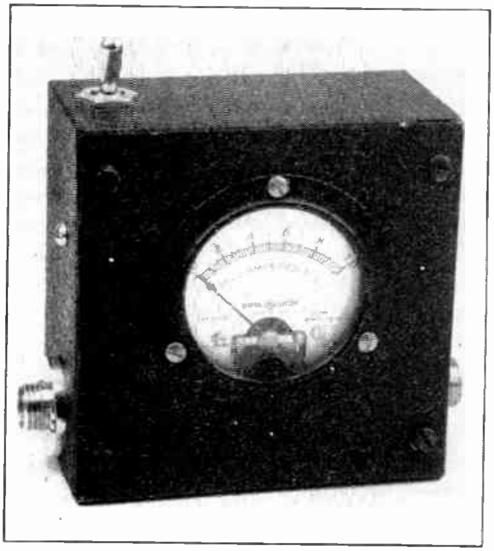


Fig. 21-49 — Resistance-bridge standing-wave indicator for coaxial lines. Input and output terminals are at the lower left and lower right, respectively. This unit, built in a 2 by 4 by 4 box, is provided with a switch so that the voltmeter can measure either the applied voltage or the bridge voltage.

ing the output connections from open or short-circuit (the reference readings) 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 as R_2 (matched as closely as possible) as a load.

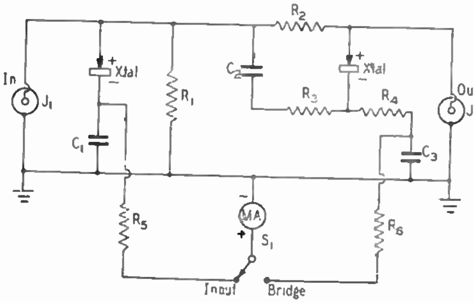


Fig. 21-50 — Resistance-bridge s.w.r. indicator for coaxial lines.

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, or microammeter.
S_1 — S.p.d.t. toggle.
Xtal — 1N51 or 1N34.

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. Because the bridge operates at a very low power level it may be necessary to couple it to a low-power driver stage rather than to the final amplifier. Alternatively, the plate voltage and excitation for the final amplifier may be reduced to the point where the power output is of the order of five watts. Then connect the test resistor to the output terminals, using leads as short as possible, and 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 loads. 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. The s.w.r., is given by

$$S.W.R. = \frac{R_L}{R_0} \text{ or } \frac{R_0}{R_L}$$

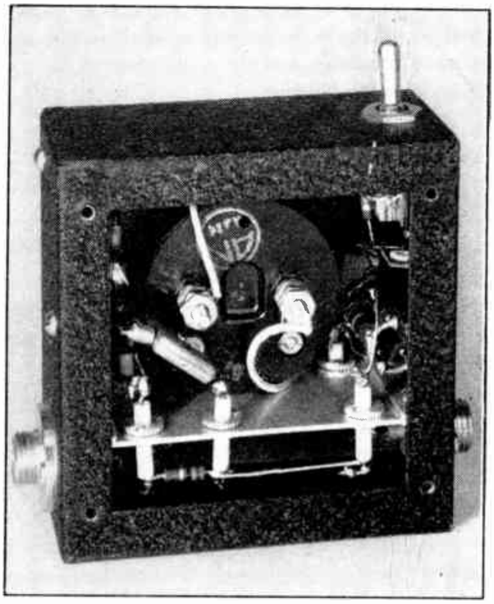


Fig. 21-51 — Inside view of the resistance-bridge s.w.r. indicator. The input terminal is at the right. An aluminum strip, the full width of the box, serves as a low-inductance ground plane for the instrument. Small ceramic through-bushings are used to insulate the "hot" line conductor and the bridge resistor, R_2 (Fig. 21-50), 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.

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 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. 21-50 the variation should not exceed about 5 per cent, and the error can be made smaller by using a low-range microammeter with a large series resistance as a voltmeter.

To use the bridge for s.w.r. measurements, connect it to the transmitter and adjust the coupling to make the meter read full scale with the bridge output terminals either open or short-circuited and S_1 in the bridge position. Check the line voltage. Connect the transmission line to be measured, readjust the transmitter coupling for the same line voltage, if necessary, and then switch S_1 to the bridge position for the s.w.r. reading, as given by the previously-determined calibration curve.

Parallel-Conductor Lines

Bridge measurements made directly on parallel-conductor lines are frequently subject to considerable error because of "antenna" currents flowing on such lines. These currents, which are either induced on the line by the field around the antenna or coupled into the line from the transmitter by stray capacitance, are in the same phase in both line wires and hence do not balance out like the true transmission-line currents. They will nevertheless actuate the bridge voltmeter, causing an indication that has no relationship to the standing-wave ratio.

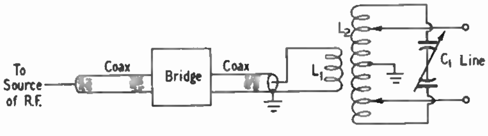


Fig. 21-52 — Circuit for using coaxial s.w.r. bridge for measurements on parallel-conductor lines. Values of circuit components are identical with those used for the similar "antenna-coupler" circuit discussed in the chapter on transmission lines.

The effect of "antenna" currents on s.w.r. measurements can be largely overcome by using a coaxial bridge and coupling it to the parallel-conductor line through a properly-designed impedance-matching circuit. A suitable circuit is given in Fig. 21-52. It closely resembles the common type of "antenna coupler", and in fact such a coupler can be used for the purpose. In the balanced tank circuit the "antenna" or parallel components on the line tend to balance out and so are not passed on to the s.w.r. bridge. It is essential that L_1 be coupled to a "cold" point on L_2 to minimize capacitive coupling, and also desirable that the center of L_2 be grounded to the chassis on which the circuit is mounted. Values should be such that L_2C_1 can be tuned to the operating frequency and that L_1 provides sufficient coupling, as described in the transmission-line chapter. The measurement procedure is as follows:

Connect a noninductive (½ or 1 watt carbon) resistor, having the same value as the characteristic impedance of the parallel-conductor line, to the "line" terminals. Apply r.f. to the bridge, adjust the taps on L_2 (keeping them equidistant from the center) and vary the capacitance of C_1 until the bridge shows a null. After the null is obtained, do not touch any of the circuit adjustments. Next, short-circuit the "line" terminals and adjust the r.f. input until the bridge voltmeter reads full scale. Remove the short-circuit and test resistor, and connect the regular transmission line. The bridge will then indicate the standing-wave ratio on the line.

The circuit requires rematching, with the test resistor, whenever the frequency is changed appreciably. It can, however, be used over a portion of an amateur band without readjustment, with negligible error.

The "Twin-Lamp"

A simple and inexpensive standing-wave indicator for 300-ohm line is shown in Fig. 21-53. 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. The power input to the line should be adjusted to make the lamp nearest the transmitter light to full brilliance. If the line is properly matched and the reflected power is very low, the lamp toward the antenna will be dark. If the s.w.r. is high, the two lamps will glow with practically equal brilliance.

The length of the piece of 300-ohm line used in the twin-lamp will depend on the transmitter power and the operating frequency. A few inches will suffice with high power and at high fre-

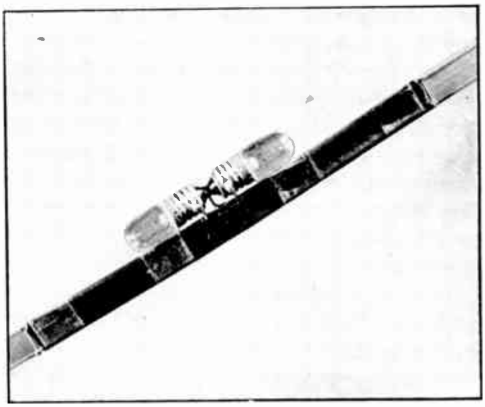


Fig. 21-53 — The "twin-lamp" standing-wave indicator.

quencies, while a foot or two may be needed with low power and at low frequencies.

In constructing the twin-lamp, 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. Remove about ¼ inch of insulation from one wire of the main transmission line at some convenient point. 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 transmission line, then solder the ends of the cut portion of the short piece to the shells of the bulbs. Figs. 21-53 and 21-54 should make the construction clear.

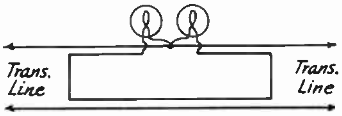


Fig. 21-54 — **Wiring diagram of the "twin-lamp" standing-wave indicator.**

Installing the twin-lamp on a line introduces a discontinuity in the line impedance which causes the s.w.r. from the twin-lamp back to the transmitter to differ from the s.w.r. existing between the antenna and twin-lamp. For this reason it is frequently desirable to remove it after s.w.r. checks have been made. It is convenient to mount the twin-lamp on a short length of line fitted to a 300-ohm plug at one end and a mating socket (crystal socket with half-inch spacing) at the other. If similar plugs and sockets are used on the transmitter and regular transmission line, the whole test unit can be inserted and taken out at will.

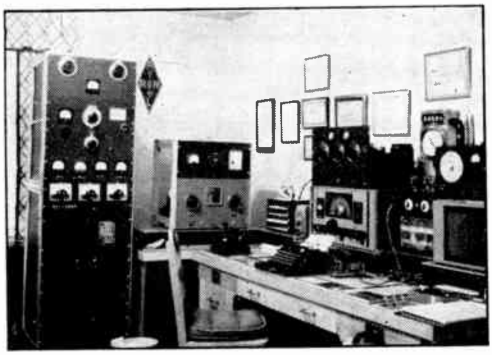
The twin-lamp will respond to "antenna" currents on the transmission line in much the same way as the bridge circuits discussed earlier. There is therefore always a possibility of error in its indications, unless it has been determined by other means that "antenna" currents are inconsequential compared with the true transmission-line current.

CHAPTER 22

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 cubby-holes provided for message forms, log book, Call Book and other papers keep the operating position neat and ready for action at any time. (W4CDA, 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.

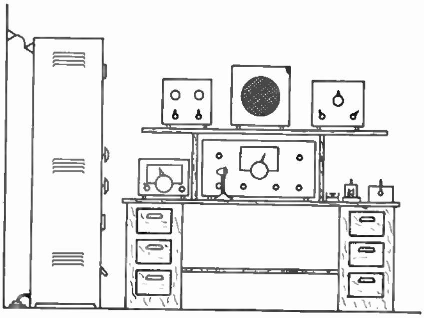


Fig. 22-1 — **In a station assembled for maximum ease in frequency or band changing, the transmitter should be located next to the operating position, as shown above. On the operating table, the receiver is in front of the operator and VFO or crystal-switching oscillator on the left. (The VFO or crystal 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.

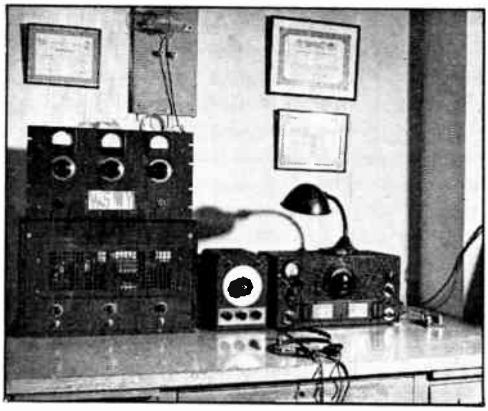
A compromise arrangement would place the VFO or crystal-switched oscillator at the operating position and the transmitter in some convenient location not adjacent to the operator. Since it is usually possible to operate over a portion of a band without retuning the transmitter stages, an operating position 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.

One of the most convenient station arrangements is to build a semi-circular operating table as shown here. All operating controls are readily available, and considerably more equipment can be grouped around the operator than when an ordinary desk is used. (W2SAI, Riverton, N. J.)

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 for 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 are not too desirable, although they are widely used in mobile and portable work.

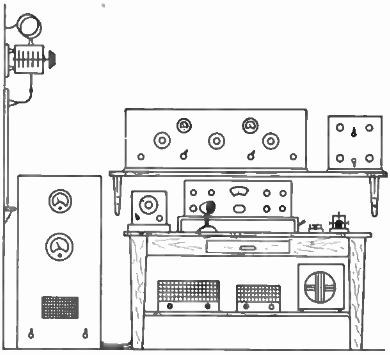
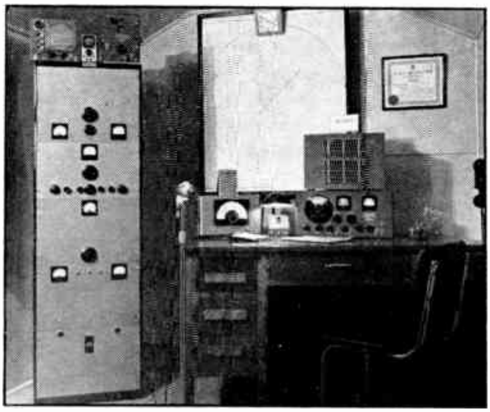


Fig. 22-2 — When little space is available for the amateur station, the equipment has to be spotted where it will fit. In the above arrangement, the transmitter, modulator and power supplies (separate units) are sandwiched in alongside the operating table and on a shelf above the table. The antenna tuning 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 illustrates how concealing all interconnecting wires and eliminating gear not necessary to communication results in an extremely neat station. (VE3AUJ, Woodstock, Ont.)

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 controlled by the switch.

If a rotary beam is used the control of the beam should be convenient to the operator. The 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 included with the 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, 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 pur-

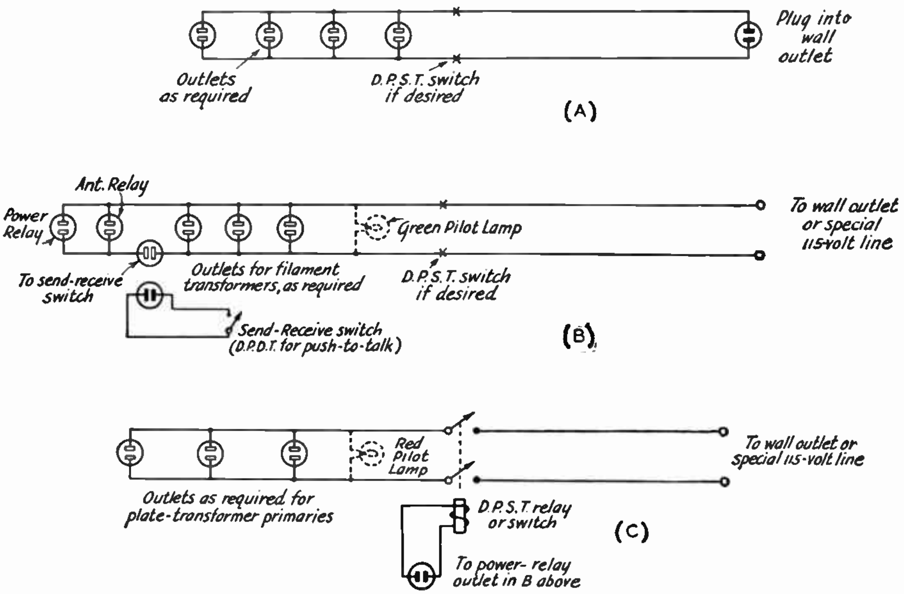


Fig. 22-3 — Power circuits for a high-power station. A shows the outlets for the receiver, monitoring equipment, speech amplifier and the like. The outlets should be mounted inconspicuously on the operating table. B shows the transmitter filament circuits and control-relay circuits, if the latter are 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.

pose, 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.

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.

It is neater and safer 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.

Interconnections

The wiring of any station will entail two or three common circuits, as shown in Fig. 22-3. 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.

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. (See chapter on Power Supplies for high-power considerations. 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 that 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 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 current ratings.

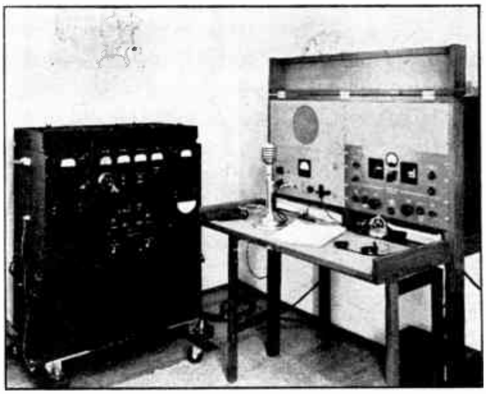
When relays are used, the send-receive switch closes the circuit to their coils, thus closing the relay contacts. The relay contacts are in the power circuit being controlled, and thus the switch handles only the relay-coil current.

● 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



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.)

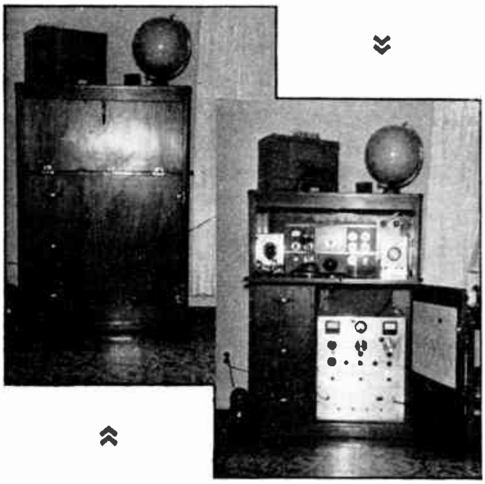
one or more of these components may leave the transmitter in a dangerous condition. The shorting stick is made by mounting a small metal hook, of wire or rod, on one end of a dry stick or bakelite rod. A piece of ignition cable or other well-insulated wire is then run from the hook on the stick to the 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. This simple device, 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÷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.



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.)

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.)

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 changeover 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 line, it simplifies the problem of "What to do with the feeders?"

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 23

BCI and TVI

It is the duty of every amateur to make sure that the operation of his station does not, because of any shortcomings in equipment, cause interference with other radio services.

However, there is a larger obligation — to eliminate interference with regular broadcasting (BCI) and television (TVI) 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 programs are broken up by amateur transmissions. 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 much interference is directly the fault of 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 interference occurs because receivers are not capable of rejecting signals far outside the frequency band to which the receiver is tuned. "Quiet hours" are not imposed unless it is shown that the interference is actually the fault of the transmitter.

● GETTING LISTENER COÖPERATION

To be successful in handling interference cases you have got to win the listener's coöperation. The first step is to earn the listener's confidence in your technical ability and to convince him of 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. The best check on this is your own AM or TV receiver. It is always convincing if you can say — and demonstrate — that you do not interfere with reception in your own home.

Don't Hide Your Identity

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

The average person will tolerate a limited amount of interference, but no one can be expected to put up with frequent and extended interruptions to programs. 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

When you interfere, it is natural for the complainant to assume that your transmitter is at fault. Explain that you do not operate on the broadcast frequencies, and the real trouble is that you and he happen to be located so close to each other. Point out that the average 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 filter or wavetrap. If the wiring of the receiver itself is picking up your signal, 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 receiver. You can then determine for yourself where the trouble is most likely to be.

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.

However, if the owner of the receiver obviously prefers to have you make the modifications, do so only with the understanding that it is purely because you are anxious to coöperate.

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 interference problems on the air, do it in a constructive way — one calculated to increase listener coöperation, not destroy it.

● RADIO-CLUB INTERFERENCE 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 amateur interference. 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 interference 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 nonlinear 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.) 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 if the interference is to be eliminated. Recommend that the actual work be done by a radio serviceman. Offer to check into the cause yourself, if he 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 it does, 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.

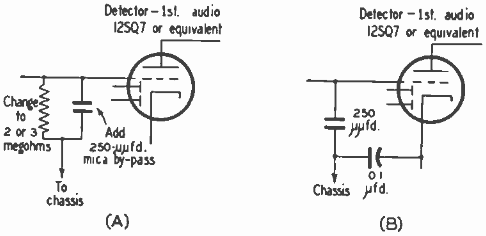


Fig. 23-1 — Two methods of eliminating r.f. from the grid of a combined detector/first-audio stage. At A, the value of the grid leak is reduced to 2 or 3 megohms, and a mica by-pass condenser is added. At B, both grid and cathode are by-passed.

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.

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. 23-1 and 23-2. Fig. 23-1A 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-µµfd. mica condenser. Fig. 23-1B is a similar method. A third method that has worked in 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. 23-2 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.

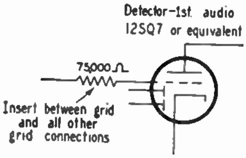


Fig. 23-2 — Using a 75,000-ohm resistor to form a low-pass filter with the tube capacitance. The resistor must be mounted at the tube pin, between the grid and all other grid connections.

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. 23-3. If interference results from operation in more than one amateur band several traps may be connected 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.

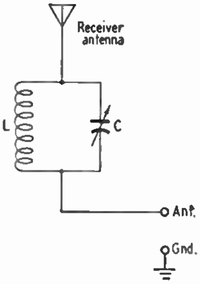


Fig. 23-3 — A simple wavetrap circuit. *L* and *C* must resonate at the frequency of the interfering signal. Suitable constants are tabulated below.

Band	*C*	*L*
3.5	140 µµfd.	16 µh., 32 turns #22, 1″ diam., 1″ long
7	100 µµfd.	6 19 #22, 1″ 1″
14	50 µµfd.	3.5 14 #18, 1″ 1″
21	35 µµfd.	2.2 12 #18, 1″ 1″
28	25 µµfd.	1.5 9 #18, 1″ 1″

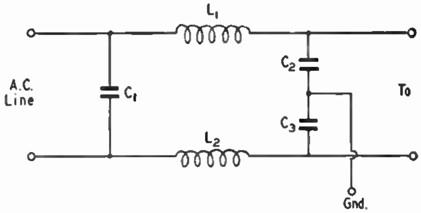


Fig. 23-4 — A.c. line filter for receivers. The values of C_1, C_2 and C_3 are not generally critical; capacitances from 0.001 to 0.01 µfd. can be used. L_1 and L_2 can be a 2-inch winding of No. 18 enameled wire on a half-inch diameter form.

A common form of a.c. line filter is shown in Fig. 23-4. 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.

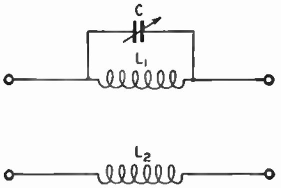


Fig. 23-5 — Resonant filter for the a.c. line. A single condenser tunes both L_1 and L_2, which are unity-coupled, one wound on top of the other. Constants for amateur bands are tabulated below.

Band	*C*	L_1 - L_2
3.5	140 + 150 (fixed)	25 t. No. 18, 1¼″ dia. × 2⅜″ long
7	140 µµfd.	18 t. No. 18, 1¼″ dia. × 2⅜″ long
14	100 µµfd.	12 t. No. 18, 1¼″ dia. × 2⅜″ long
21	50 µµfd.	10 t. No. 18, 1¼″ dia. × 2⅜″ long
16	25 µµfd.	9 t. No. 18, 1½″ dia. × 2⅜″ long

D.c.c. wire is recommended for all coils.

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. 23-5 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 shield box, to prevent pick-up in the filter and to make it less conspicuous.

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 in principle 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. 23-6. 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 sixth — 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.

It is very difficult to secure adequate harmonic suppression, especially with transmitters operating at 14 Mc. and higher, unless these three phases are attacked separately and in the order given above. It is impossible to build a transmitter that will not generate *some* harmonics, but it is obviously advantageous to reduce their strength, by circuit design, by as large a factor as possible before attempting to prevent them from being radiated. Second, harmonic radiation from the transmitter itself or from its associated wiring obviously will cause interference just as readily as radiation from the antenna; so measures taken to prevent harmonics from reaching the antenna will not reduce TVI if the transmitter itself is radiating harmonics. But once it has been found that the transmitter itself is free from harmonic radiation, devices for preventing harmonics from reaching the antenna can be expected to produce results.

There is no magic "gimmick" that will eliminate TVI caused by harmonics. The problem has to be worked on one step at a time.

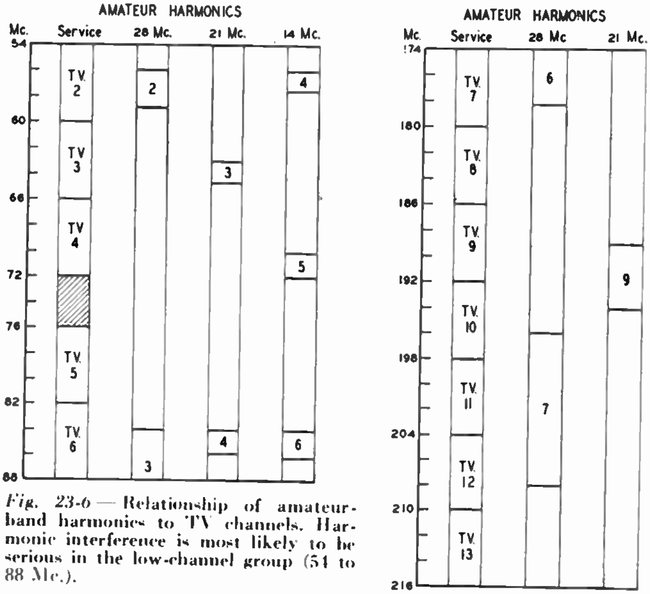


Fig. 23-6 — **Relationship of amateur-band harmonics to TV channels. Harmonic interference is most likely to be serious in the low-channel group (54 to 88 Mc.).**

REDUCING HARMONIC GENERATION

Reasonably-efficient operation of r.f. power amplifiers always is accompanied by harmonic generation, and in the case of frequency multipliers the harmonic output at a particular frequency is deliberately accentuated. From the standpoint of TVI reduction, good judgment calls for operating all frequency-multiplier stages at a very low power level — receiving tubes and plate voltages not exceeding 250 or 300. When the final output frequency is reached, it is highly desirable to use as few stages as possible in reaching the output power level, and to use tubes that require a minimum of driving power. The smaller the number of stages operating at appreciable power levels, the smaller the number of points where damaging harmonics can be generated.

Circuit Design and Layout

Harmonic currents of considerable amplitude flow in both the grid and plate circuits of r.f. power amplifiers. They will do relatively little harm if they can be effectively by-passed to the cathode of the tube, but this is frequently difficult to do. Fig. 23-7A shows the paths followed by harmonic currents in an amplifier circuit; because of the high reactance of the tank coil there is little harmonic current in it, so the harmonic currents simply flow through the tank condenser, the plate (or grid) blocking condenser, and the tube capacitances. The lengths of the leads forming these paths is of great importance, since the inductance in this circuit will resonate with the tube capacitance at some frequency in the v.h.f. range (the tank and blocking condensers are so large compared with the tube capacitance that they have little effect on the resonant frequency). If such a resonance happens to occur at or near the same frequency as one of the transmitter harmonics, the effect is just the same as though a harmonic tank circuit had been deliberately introduced; the harmonic at that frequency will be tremendously increased in amplitude.

Such resonances are unavoidable, but by keeping the path from plate to cathode and from grid to cathode as short as is physically possible, the resonant frequency usually can be raised above 100 Mc. in amplifiers of medium power. This puts it between the two groups of television channels. Except in very low power miniature-tube transmitters, it is usually not feasible to raise the resonance above 216 Mc.

Where physically-short return paths from plate or grid to cathode are difficult because of the shape and size of tubes and tank condensers, the arrangement shown in Fig. 23-7B is frequently helpful. Condensers C_5 and C_6 should be of the vacuum or tubular type and should be mounted as close as possible to the tube connections. They form resonant circuits in themselves with the tube capacitance, but generally at a sufficiently high frequency so that no harm is done. At lower frequencies than this self-resonance, they effectively add to the tube capacitance and thus tune the inductance of the leads through the regular tank and blocking condensers to a considerably lower frequency than the tube alone. The resonance therefore can be shifted to a frequency *below* 54 Mc. and again is outside the TV range.

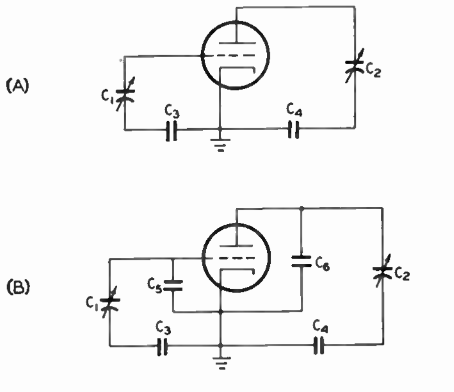


Fig. 23-7 — (A) A v.h.f. resonant circuit is formed by the tube capacitance and the leads through the tank and blocking condensers. Regular tank coils are not shown, since they have little effect on such resonances. (B) Using low-inductance condensers shunting the tube elements to lower the resonance point below the TV channels. C_5 and C_6 usually are 15 to 50 μμfd. and either of vacuum or tubular construction.

This method is most useful at 3.5 and 7 Mc. It increases the tank capacitance to the point where there may be very little tank coil left, when the transmitter is used on 28 Mc., unless the leads are eliminated by using the shunting condenser as the tank condenser and adjusting the tank coil inductance to resonate, no regular tank condenser being used.

It is easier to place grid-circuit v.h.f. resonances where they will do no harm if the amplifier is link-coupled to the driver stage, since this generally permits shorter leads and more favorable conditions for by-passing the harmonics than is the case with capacitive coupling. Link coupling also reduces the coupling between the driver and amplifier at harmonic frequencies, thus preventing driver harmonics from being amplified.

The inductance of leads from the tube to the tank condenser can be reduced not only by shortening but by using flat strip instead of wire conductors. It is also better to use the chassis as the return from the blocking condenser to cathode, since a chassis path will have less inductance than almost any other form of connection.

The v.h.f. resonance points in amplifier tank circuits can be found by coupling a grid-dip meter covering the 50–250 Mc. range to the grid and plate leads. If a resonance is found in or near a TV channel, methods such as those described above should be used to move it well out of the TV range. The grid-dip meter also should be used to check for v.h.f. resonances in the tank coils, because coils made for 14 Mc. and below usually will show such resonances. If a resonance falls in a TV channel that is in use in the locality, changing the number of turns will move it to a frequency where it will not be troublesome.

In most r.f. amplifiers the cathode connection of the tube is below chassis while the plate (and sometimes the grid) connection frequently is above. In such a case the blocking condenser should be mounted *below* chassis. If the ground return is made to the top, the r.f. current has to flow over the top and either through a good-sized hole or else entirely over the chassis surface before it reaches the cathode. This condition is highly undesirable not only because of v.h.f. resonances but because such chassis currents frequently cause instability in the amplifier. If the by-pass condenser is mounted above, it should be connected to the cathode by means of an insulated lead running through the chassis by the shortest possible path.

Operating Conditions

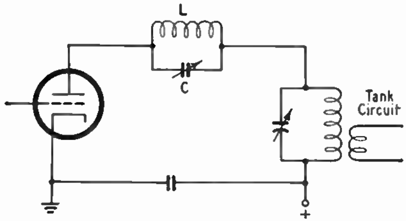
High values of grid bias and grid current increase the harmonic content of the r.f. currents in both the grid and plate circuits. All tubes in the transmitter, and particularly those operating at appreciable power levels, should be driven no harder than is necessary to give reasonably efficient operation and satisfactory linearity, if modulated. Generally, it is unnecessary to go very far beyond cut-off bias, and the grid current should be kept to the minimum that gives satisfactory operation.

For equal operating conditions, there is little or no difference between single-ended and push-pull amplifiers in respect to harmonic generation. Push-pull amplifiers are frequently trouble-makers on even harmonics because with such amplifiers the even-harmonic voltages are in phase at the ends of the tank circuit and hence appear with equal amplitude across the whole tank coil, if the center of the coil is not grounded. Under such circumstances the even harmonics can be coupled to the output circuit through stray capacitance between the tank and coupling coils. This does not occur in a single-ended amplifier if the coupling coil is placed at the cold end of the tank.

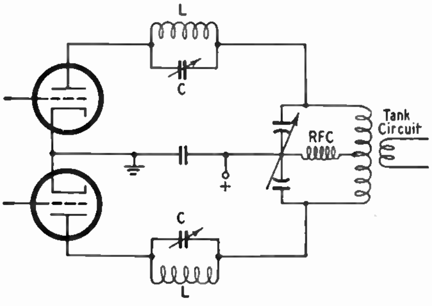
Harmonic Traps

If a harmonic in only one TV channel is particularly bothersome — frequently the case when the transmitter operates on 28 Mc. — its amplitude can be reduced by a very considerable factor if a trap tuned to the harmonic frequency is installed in the plate lead as shown in Fig. 23-8. At the harmonic frequency the trap represents a very high impedance and hence reduces the amplitude of the harmonic current flowing through the tank circuit. In the push-pull circuit both traps have the same constants. The *L/C* ratio is not critical but a high-*C* circuit usually will have least effect on the performance of the plate circuit at the normal operating frequency.

Since there is a considerable harmonic voltage built up across the trap, there may be radiation from the trap unless the transmitter is well shielded. The traps should be placed so that there is no coupling between them and the amplifier tank circuit.



SINGLE-ENDED



PUSH-PULL

Fig. 23-8 — Harmonic traps in an amplifier plate circuit. *L* and *C* should resonate at the frequency of the harmonic to be suppressed. *C* may be a 25- to 50-$\mu\mu$fd. midget, and *L* usually consists of 3 to 6 turns about ½ inch in diameter. The inductance should be adjusted so that the trap resonates at about half capacity of *C* before being installed in the transmitter. It may be checked with a grid-dip meter. When in place, it is adjusted for minimum interference to the TV picture.

A trap is a highly-selective device and so is useful only over a small range of frequencies. A second- or third-harmonic trap on a 28-Mc. tank circuit usually will not be effective over more than 50 kc. or so at the fundamental frequency, depending on how serious the interference is without the trap. Because they are critical of adjustment, it is better to prevent TVI by other means, if possible, and use traps only as a last resort.

PREVENTING RADIATION FROM THE TRANSMITTER

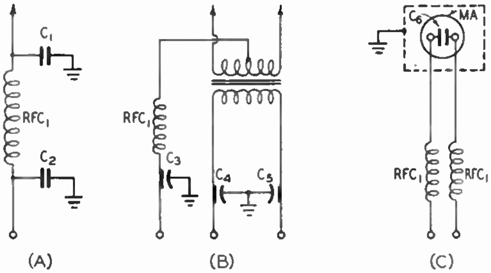
The extent to which harmonic interference will be caused by transmitter radiation depends on the operating frequency, the transmitter power level, the strength of the television signal, and the distance between the transmitter and TV receiver, as well as on the strength of the harmonics generated in the transmitter. Transmitter radiation can be a very serious problem if the TV signal is marginal or below, the TV receiver and amateur transmitter are close together, and the transmitter is operated with high power on 28 Mc.

Transmitter radiation can be prevented by shielding the r.f. chassis. A metal enclosure is not necessarily a shield. A shield will not be good, electrically, unless all its joints make good connections along their entire length. A slit or crack at a joint will let out a surprising amount of r.f. energy. Ventilating louvers and large holes such as those used for mounting meters will do the same. On the other hand, small holes do not impair the shielding very much, hence ventilating holes may be used if they are small — not over ¼ inch in diameter — and wire screening of the type used for fly screens also is satisfactory. It is unfortunate that conventional metal cabinets, with their doors, painted surfaces where joined, and ventilating louvers are practically useless as shields. They can be made effective by covering louvers with screening, and by scraping and bonding all joints to secure a good electrical connection.

Lead Treatment

Even very good shielding can be made completely useless when connections are run from external power supplies and other equipment to the circuits inside the shield. Every conductor so introduced into the shielding forms a path for the escape of r.f., which is then radiated by the connecting wires. Hence a step that is essential in every case, and more important than the shielding itself in most, is to prevent harmonic currents from flowing on the leads leaving the shielded enclosure.

Harmonic currents always flow on the d.c. or a.c. leads connecting to the tube circuits. A very effective means of preventing such currents from being coupled into other wiring, and one that provides desirable by-passing as well, is to use shielded wire for all such leads, maintaining the shielding from the point where the lead connects to the tube or r.f. circuit right through to the point where it is about to leave the chassis. The shield braid should be grounded to the chassis at both ends and at frequent intervals along the path. Where the lead comes to the connection block through which it leaves the chassis it is frequently necessary to install r.f. filters as well. Typical filter circuits are shown in Fig. 23-9. The "Hypass" type of condenser is most effective at v.h.f., but mica condensers will serve when installed so that there is essentially no lead length between ground (chassis) and the circuit to which the ungrounded side of the condenser is connected. It is generally best to use low (not more than 500 $\mu\mu$fd.) values of capacitance when mica by-passes are used, and to employ the smallest physical size of condenser that is rated for the voltage on the circuit.



(A) (B) (C)

Fig. 23-9 — 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 may be 470-$\mu\mu$fd. mica condensers. Feed-through type condensers (Sprague "Hypass") are recommended for C_3, C_4 and C_5 because it is difficult to build satisfactory low-resistance chokes for filament circuits. 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 or equivalent).

Filters at the connection block preferably should be shielded from each other and from the other components under the chassis. However, if the coils are physically small — they usually consist of a winding of an inch or so on a ¼-inch diameter form — and are mounted well away from other components, it is possible to dispense with individual shielding of the filters.

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. 23-9C 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, to include the resistance of the leads and chokes.

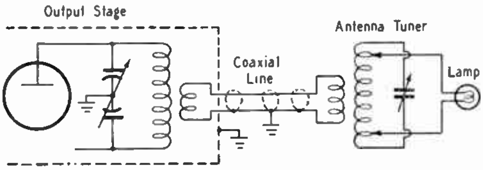


Fig. 23-10 — Recommended dummy-antenna circuit for checking harmonic radiation from the transmitter and leads. The matching circuit helps prevent harmonics in the output of the transmitter from flowing back over the transmitter itself, as might occur if the lamp load is simply connected to the output coil of the final amplifier. See transmission-line chapter for details of the matching circuit. Tuning must be adjusted by cut-and-try, since the bridge method described in the transmission-line chapter will not work with lamp loads because of the change in resistance when the lamps are hot.

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. Where the wiring crosses or runs parallel, the shields should be spot-soldered together and connected to the chassis.

Checking Transmitter Radiation

The simplest instrument for checking harmonic radiation from the transmitter chassis and the connecting leads is a crystal-detector wavemeter covering the TV range (see chapter on measurements). In most cases it will be sufficient to cover 50 to 100 Mc., since harmonics falling in the 176-216 Mc. range will be of low amplitude from transmitters working on 28 Mc. and below, providing the principles of reducing harmonic generation have been followed. In the case of a 50-Mc. transmitter, the methods discussed above are used, but the only troublesome harmonic is the fourth, falling in the neighborhood of 200 Mc. In such case the wavemeter should cover 200 Mc.

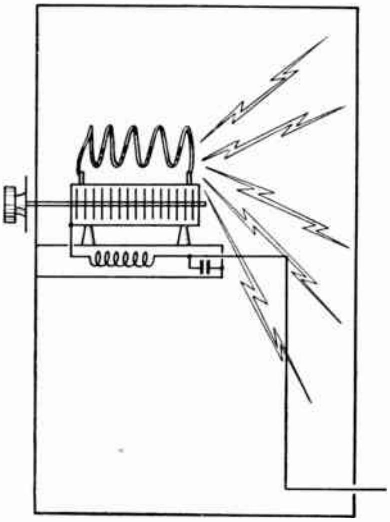


Fig. 23-11 — A metal cabinet can be an adequate shield, but there will still be radiation if the leads inside can pick up r.f. from the transmitting circuits.

Checking should be done with the transmitter delivering full power into a dummy antenna, preferably through a coax-matching circuit such as is shown in Fig. 23-10. The lamp used as the dummy antenna should have a power rating equal to the power output of the transmitter; a combination of lamps in series or parallel may be used instead of a single lamp. By exploring the vicinity of the transmitter shield with the wavemeter, tuning over the TV range, the direct radiation may be checked and the weak spots in the shielding located. The presence of harmonic currents on the external leads may be detected by wrapping a turn or two of the lead around the wavemeter coil. If a harmonic can be detected at all in a lead, a different filter arrangement should be tried until the wavemeter no longer indicates its presence. When the wavemeter shows no harmonics to be either radiated from the shielding or on the leads, the transmitter will be satisfactory in most localities. If the TV signal is very weak, a more sensitive indicator such as the regenerative wavemeter shown in the measurements chapter should be used.

Proper shielding of the transmitter requires that the r.f. circuits be shielded entirely from the external connecting leads. A situation such as is shown in Fig. 23-11, where the leads in the r.f. chassis have been shielded and properly filtered but the chassis is mounted in a large shield, simply invites the harmonic currents to travel over the chassis and on out over the leads *outside* the chassis. The shielding about the r.f. circuits should make complete contact with the chassis on which the parts are mounted.

PREVENTING HARMONICS FROM REACHING THE ANTENNA

The third and last step in reducing harmonic TVI is to keep the harmonics generated in the final stage from traveling over the transmission line to the antenna. It is seldom worthwhile even to attempt this until the radiation from the transmitter and its connecting leads has been reduced to the point where, with the transmitter delivering full power into a dummy antenna, it has been determined both by the wavemeter check and by actual testing with a television receiver that the radiation is below the level that can cause interference. If the dummy antenna test shows enough radiation to be seen in a TV picture, it is a practical certainty that harmonics will be coupled to the antenna system no matter what preventive measures are taken.

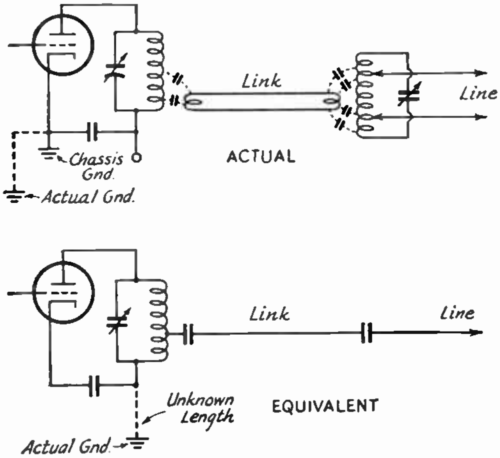


Fig. 23-12 — The stray capacitive coupling between coils in the upper circuit leads to the equivalent circuit shown below.

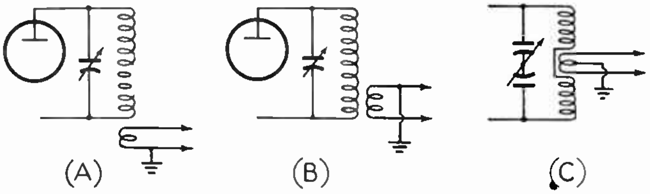


Fig. 23-13 — Methods of coupling and grounding link circuits to reduce energy transfer through stray capacitance.

In inductively-coupled output systems, some harmonic energy will be transferred from the final amplifier through the mutual inductance between the tank coil and the output coupling coil. Harmonics transferred in this way are not too hard to handle, and can be greatly reduced by providing sufficient selectivity between the final tank and the transmission line. A good deal of selectivity, amounting to 20 to 30 db. reduction of the second harmonic and much higher reduction of higher-order harmonics, is furnished by a matching circuit of the type shown in Fig. 23-10 and described in the chapter on transmission lines. An "antenna coupler" is therefore a worthwhile addition to the transmitter.

Capacitive Coupling

Harmonics transferred from the tank by stray capacitance are not suppressed by an antenna coupler to the same extent as those transferred by pure inductive coupling. The upper drawing in Fig. 23-12 shows the 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. Energy coupled through these capacitances travels over the link circuit and the transmission line as though these were merely single conductors. The tuned circuits simply act as masses of metal and offer no selectivity at all for capacity-coupled energy. Although the actual capacitances are small, they offer a very good coupling medium for frequencies in the v.h.f. range.

Capacitive coupling can be reduced by coupling to a "cold" point on the tank coil — the end connected to ground or cathode in a single-ended stage. In push-pull circuits having a split-stator condenser with the rotor grounded for r.f., all parts of the tank coil are "hot" at even harmonics, but the center of the coil is "cold" at the fundamental and odd harmonics. If the center of the tank coil, rather than the rotor of the tank condenser, is grounded through a by-pass condenser the center of the coil is "cold" at all fre-

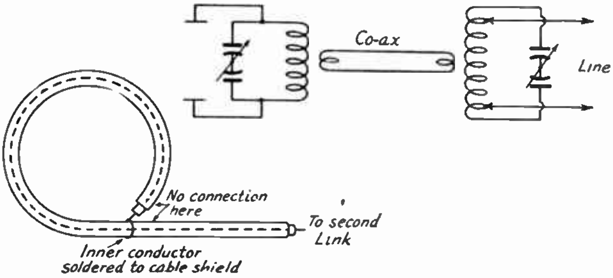


Fig. 23-14 — **A shielded link coil constructed from coaxial cable. The smaller sizes of cable such as RG-59/U are most convenient, except when the coils have a diameter of 3 inches or more. For larger coils, RG-8/U or RG-11/U can be used.**

quencies, but this arrangement is not very desirable because it causes the harmonic currents to flow through the coil rather than the tank condenser and thus increases the harmonic transfer by pure inductive coupling.

With either single-ended or balanced tank circuits the coupling coil should be grounded to the chassis by a short, direct connection as shown in Fig. 23-13. If the coil feeds a balanced line or link, it is preferable to ground its center, but if it feeds a coax line or link one side may be grounded. Coaxial output is much preferable to balanced output, from the standpoint of harmonic reduction.

At high frequencies — 28 and possibly 14 Mc. — capacitive coupling can be greatly reduced by using a shielded coupling coil as shown in Fig. 23-14. The inner conductor of a length of coaxial cable is used to form a one-turn coupling coil. The outer conductor serves as an open-circuited shield around the turn, the shield being grounded to the chassis. The shielding has no effect on the inductive coupling. Because this construction is suitable only for one turn, the coil is not well adapted for use on the lower frequencies where many turns are required for good coupling. Shielded coupling coils having a larger number of turns are available commercially. A shielded coil is particularly useful with push-pull amplifiers when the suppression of even harmonics is important.

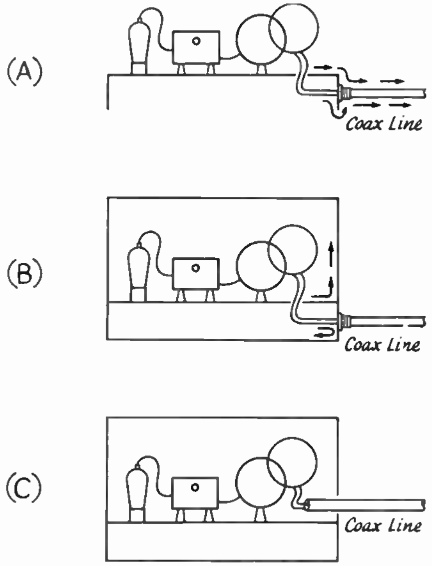


Fig. 23-15 — **Right (B) and wrong (A and C) ways to connect a coaxial line to the transmitter. In either A or C, harmonic energy coupled by stray capacitance to the outside of the cable will flow without hindrance to the antenna system. In B the energy cannot leave the shield and hence can flow out only through, not over, the cable.**

A shielded coupling coil or coaxial output will not prevent stray capacitive coupling to the antenna if harmonic currents can flow over the *outside* of the coax line. In Fig. 23-15, the arrangement at either A or C will allow r.f. to flow over the outside of the cable to the antenna system. The proper way to use coaxial cable is shield the transmitter completely, as shown at B, and make sure that the outer conductor of the cable is a continuation of the transmitter shielding. This prevents r.f. inside the transmitter from getting out by any path except the *inside* of the cable. Harmonics flowing *through* a coax line can be stopped from reaching the antenna system by an antenna coupler or by a low-pass filter installed in the line.

Low-Pass Filters

A low-pass filter properly installed in a coaxial line, feeding either a matching circuit (antenna coupler), or feeding the antenna directly, will provide very great attenuation of harmonics. The coax-coupled matching-circuit arrangement is highly recommended when the main transmission line is of the parallel-conductor type.

A properly-designed low-pass filter will not introduce appreciable power loss at the fundamental frequency if the coaxial line in which it is inserted is terminated so that the s.w.r. is low. The s.w.r. can easily be measured by means of a simple bridge as described in the chapters on measurements and transmission lines. Such a filter has the property of passing without loss all frequencies below its "cut-off" frequency, but simultaneously has large attenuation for all frequencies above the cut-off frequency. Space does not permit a complete description here, but detailed design information can be found in a series of articles in *QST* (Grammer, "Eliminating TVI With Low-Pass Filters", *QST*, in three parts, February, March, and April, 1950).

A relatively simple low-pass filter of the *m*-derived type is shown in Figs. 23-16 to 23-18, inclusive. 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. 23-17 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. 23-18, 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 filter can be adjusted by short-circuiting point *A* to the common ground plate (use the shortest possible connection) and setting 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. 23-18, 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 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.

The harmonic attenuation provided by filters of the type shown will, with careful adjustment, be adequate for areas in which the television signal is of good strength, for operation on all bands from 30 Mc. down. In weak-signal regions, more attenuation may be needed when the transmitter is operating on 28 Mc. Information on such filters may be found in the *QST* article cited above.

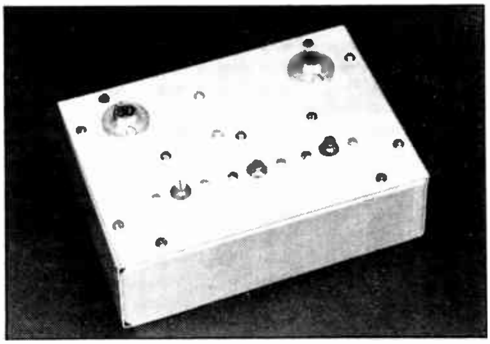


Fig. 23-16 — Television-frequency harmonic filter, for use with coax cable. All parts are mounted on a 5 × 7-inch piece of aluminum, mounted with sheet-metal screws in a 5 by 7 by 2 aluminum chassis which serves as a shield.

Filter Installation

In order to give the harmonic attenuation of which it is capable, a low-pass filter must be installed in such a way that *all* the output of the transmitter flows through it. If harmonic currents

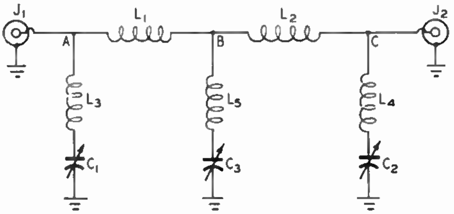


Fig. 23-17 — Circuit diagram of the harmonic filter. It provides two high-attenuation points which may be placed in television channels employed in the locality in which the filter is to be used.

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, ½-inch inside diameter, length ⅝ inch.
- L_3, L_4 — 4 turns No. 12, ½-inch inside diameter, length 9/16 inch.
- L_5 — 1 turn No. 12, ½-inch inside diameter, length ⅜ inch

For 50-ohm cable, maximum rejection in Channels 3 and 6:
- C_1, C_2 — 38 μμfd.
- C_3 — 99 μμfd.
- L_1, L_2 — 5 turns No. 12, ½-inch inside diameter, length ⅞ inch.
- L_3, L_4 — 4 turns No. 12, ½-inch inside diameter, length 11/16 inch.
- L_5 — 1 turn No. 12, ⅜-inch inside diameter, length ½ 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, ½-inch inside diameter, length ¾ inch.
- L_3, L_4 — 6 turns No. 12, ½-inch inside diameter, length 11/16 inch.
- L_5 — 3 turns No. 12, ⅜-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, ½-inch inside diameter, length 13/16 inch.
- L_3, L_4 — 6 turns No. 12, ½-inch inside diameter, length ¾ inch.
- L_5 — 2 turns No. 12, ⅜-inch inside diameter, length ¾ inch.

Coil lengths in all cases measured between centers of wire at ends.

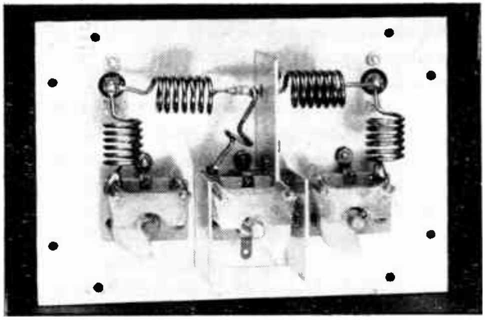


Fig. 23-18 — Construction of the harmonic filter. Dimensions should be followed fairly closely for optimum results. The center-to-center distance between the coax connectors is 4½ inches. Mounting centers of the variable condensers are on a line 2¼ inches below and parallel to a line through the centers of the coax fittings.

are permitted to flow on the outside of the connecting coaxial cables, they will simply flow over the filter and on through to the antenna, and the filter does not have an opportunity to stop them. That is why it is so important to reduce the radiation from the transmitter and its leads to negligible proportions.

Fig. 23-19 shows the proper way to install a filter between a shielded transmitter and a matching circuit. Note that the coax between the transmitter and filter, together with the shields about the transmitter and filter, forms a continuous shield to keep all the r.f. inside. It is thus forced to flow through the filter and the harmonics are attenuated. If there is no harmonic energy left after passing through the filter, shielding from that point on is not necessary; consequently, the matching circuit or antenna coupler does not need to be shielded. However, the chassis arrangement shown in Fig. 23-19 is desirable because it will tend to prevent fundamental-frequency energy from flowing from the matching circuit back over the transmitter; this helps eliminate feed-back troubles in audio systems.

If the antenna is driven through coaxial line the matching circuit shown in Fig. 23-19 may be omitted. In that case the line goes directly from the filter to the antenna.

When a filter does not seem to give the harmonic attenuation of which it should be capable, the probable reason is that harmonics are bypassing it because of improper installation and inadequate transmitter shielding, including lead filtering. However, there are occasionally cases where the circuits formed by the connecting cables and the apparatus to which they connect become resonant at a harmonic frequency. This greatly increases the harmonic output at that frequency and the overall attenuation suffers. Such troubles can be completely overcome by substituting a slightly different cable length. The most critical length is that connecting the transmitter to the filter. Checking with a grid-dip meter at the final amplifier output coil usually will show whether an unfavorable resonance of this type exists.

Summary

The methods of harmonic elimination outlined in this section have been proved beyond doubt to be effective even under highly unfavorable conditions. It must be emphasized once more, however, that the problem must be solved one step at a time, and the procedure must be in logical order. It cannot be done properly without a few items of simple equipment. These are:

1) A crystal-detector wavemeter covering the TV bands.

2) A grid-dip meter covering the TV bands.

3) A dummy antenna.

The procedure may be summarized as follows:

1) Take a critical look at the transmitter on the basis of the design considerations outlined under "Reducing Harmonic Generation".

2) Check all circuits, particularly those connected with the final amplifier, with the grid-dip meter to determine whether there are any resonances in the TV bands. If so, rearrange the circuits so the resonances are moved out of the critical frequency region.

3) Connect the transmitter to the dummy antenna and check for the presence of harmonics on leads and around the transmitter enclosure. Seal off the weak spots in the shielding and filter the leads until the wavemeter shows no indication at any harmonic frequency.

4) At this stage, check for interference on a nearby TV receiver if possible, still using the dummy antenna. If there is interference, use a more sensitive harmonic indicator (measurements chapter) and work for zero harmonic indication.

5) When the transmitter is completely clean on the dummy antenna, connect it to the regular antenna and check for interference on the TV receiver. If the interference is not bad, an antenna coupler or matching circuit installed as previously described should clear it up. Alternatively, a low-pass filter may be used. If neither the antenna coupler make any difference in the interference, the evidence is strong that the interference, at least in part, is being caused by receiver overloading because of the strong fundamental-frequency field about the TV antenna and receiver. (See later section for identification of fundamental-frequency interference.) A coupler and/or filter, installed as described above, will invariably make a difference in the interfer-

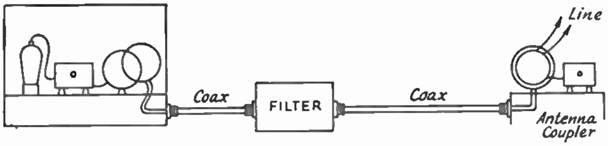


Fig. 23-19 — The proper method of installing a low-pass filter between the transmitter and antenna coupler or matching circuit. If the antenna is fed through coax the matching circuit may be omitted but the same construction should be used between the transmitter and filter. The filter should be thoroughly shielded.

ence if the interference is caused by harmonics alone.

6) If there is still interference after installing the coupler and/or filter, and the evidence shows that it is probably caused by a harmonic, more attenuation is needed. A more elaborate filter may be necessary. However, it is well at this stage to assume that part of the interference may be caused by receiver overloading, and take steps to alleviate such a condition before trying highly-elaborate harmonic-suppression methods on the transmitter.

● TV RECEIVER DEFICIENCIES

When a television receiver is quite close to the transmitter, the intense r.f. signal from the transmitter's fundamental may overload one or more of the receiver circuits to produce spurious responses that cause interference. "Fundamental" interference usually can be identified by the fact that the interference is of about the same intensity on all TV channels, including those that have no harmonic relationship to the transmitting frequency. It is a simple matter to determine, from the chart of Fig. 23-6 or by calculation, whether or not a transmitter harmonic will fall in a particular TV channel.

Assuming that the measures described earlier to reduce harmonic radiation have been taken, the presence of interference on channels that are not harmonically related to the transmitting frequency, along with interference in those that are, is a good indication that at least part of the TVI is caused by some action taking place in the receiver. The most likely possibility is that the first stage, or first few stages, are simply being overloaded. In the case of 28-Mc. operation, it is also possible that the amateur signal is getting into the picture i.f. amplifier because of insufficient i.f. rejection in the receiver.

In either case, the interference can be eliminated if the fundamental signal strength can be reduced to a level that the receiver can handle. The most satisfactory device for this purpose is a high-pass filter having a cut-off frequency between 30 and 50 Mc., installed at the antenna terminals of the receiver. Circuits that have proved effective are shown in Figs. 23-20 and 23-21. Fig. 23-21 has one more section than the filters of Fig. 23-20 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. 23-21 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.

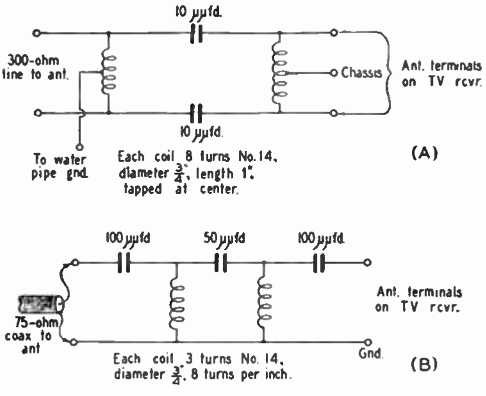


Fig. 23-20 — High-pass filters for installation at the TV receiver antenna terminals. A — balanced filter for 300-ohm line, B — for 75-ohm coaxial line. *Important:* Do not use a direct ground on an a.c.-d.c. chassis. Ground through a 0.001-µfd. mica condenser.

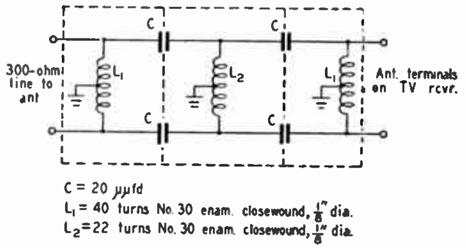


Fig. 23-21 — Another type of high-pass filter for 300-ohm line. The coils may be wound on ⅛-inch diameter plastic knitting needles. *Important:* Do not use a direct ground on an a.c.-d.c. chassis. Ground through a 0.001-µfd. mica condenser.

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. 23-3. 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.

If the fundamental signal is getting into the receiver by way of the line cord a line filter such as that shown in Fig. 23-4 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.

Antenna Installation

Many television receivers will respond strongly 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 for the receiver 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 overloading, as well as harmonic pick-up, to a level that does not interfere with reception.

CHAPTER 24

Construction Practices

● 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.

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.
Socket punches, esp. 1⅛" and 1¼".

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.

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.
Copper braid for shielding wires.

Machine screws, nuts, washers, soldering lugs, etc., are most reasonably purchased in quantities of a gross.

● 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. Aluminum is to be preferred to steel, not only because it is a superior shielding material, but because it is much easier to work and to provide good chasis contacts.

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.

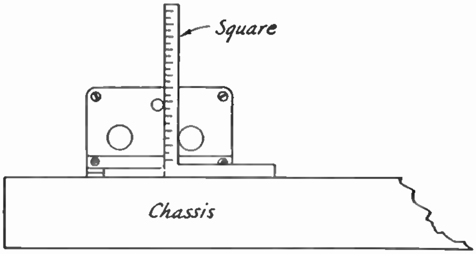


Fig. 24-1 — Method of measuring the heights of condenser shafts, etc. If the square is adjustable, the end of the scale should be set flush with the face of the head.

TABLE 24-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.

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 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. The small holes for socket-mounting screws are best located and punched, using the socket itself as a template, after the main center hole has been cut.

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 of the chassis should be transferred to the panel, by once again fastening the panel to the chassis and marking it from the rear.

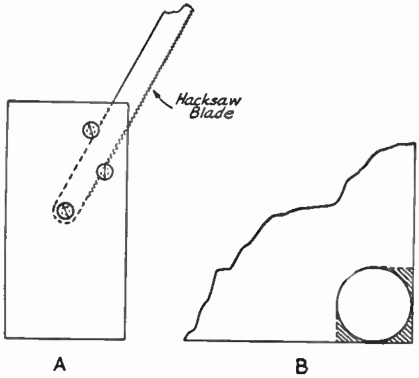


Fig. 24-2 — To cut rectangular holes in a chassis corner, holes may be filed out as shown in the shaded portion of B, making it possible to start the hack-saw blade along the cutting line. A shows how a single-ended handle may be constructed for a hack-saw blade.

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. 24-1. 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. Holes for terminals etc., in the rear edge of the chassis should be marked and drilled at the same time that they are done for the top.

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. 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. 24-2, 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.

Placement of Components

While a requirement of compactness in certain equipment may preclude best practices in the placement of parts, there are a few basic rules that should be followed whenever the necessary space can be provided. They are important from the considerations of losses and stability.

Coils should be mounted so that they are spaced at least their diameter from shielding or other large masses of metal or dielectric. This applies particularly to the ends of coils.

It should be remembered that mounting a variable condenser with its stator plates close to the chassis will increase the minimum capacitance appreciably. This may be of considerable importance when, for instance, it is desired to design a tank circuit that will have sufficiently low capacitance for 28 Mc. when using a large tank condenser that will also be suitable for 3.5 Mc. For this reason, it may be preferable to invert the condenser so that its rotor plates, rather than the stator plates, are close to the chassis when the condenser is near minimum capacitance.

Resistors and by-pass condensers may be mounted flat against the chassis, but coupling or plate-blocking condensers should be well spaced from the chassis to reduce capacitance to ground. The characteristics of r.f. chokes also may be affected by mounting too close to the chassis.

Screen and filament or cathode by-pass condensers should be mounted on the tube socket with the shortest possible leads. While the length of leads between the tank coil and the tank condenser are of less importance, it should be an objective in design to arrange the components so that the paths from the rotors of grid and plate tank condensers back to cathode or filament are as short as possible. These two paths should be independent and of heavy conductor to keep the inductance low. It is usually considered preferable not to use the chassis for these return paths.

● 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 nonmetal 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.

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

Bends may be made similarly. The sheet should be scratched on both sides, but not so deeply as to cause it to break.

Finishing Aluminum

Aluminum chassis, panels and parts may be given a sheen finish by treating them in a caustic bath. An enamelled container, such as a dishpan or infant's bathtub, should be used for the solution. Dissolve ordinary household lye in cold water in a proportion of ¼ to ½ can of lye per gallon of water. The stronger solution will do the job more rapidly. Stir the solution with a stick of wood until the lye crystals are complete dissolved. Be very careful to avoid any skin contact with the solution. It is also harmful to clothing. Sufficient solution should be prepared to cover the piece completely. When the aluminum is immersed, a very pronounced bubbling takes place and ventilation should be provided to disperse the escaping gas. A half hour to two hours in the solution should be sufficient, depending upon the strength of the solution and the desired surface.

Remove the aluminum from the solution with sticks and thereafter handle with cotton gloves until after the piece has been rinsed thoroughly in cold water while swabbing with a rag to remove the black deposit. If any black stains result, reimmerse the piece for another minute or two and rinse again. (See May, 1950 *QST* for a method of coloring and anodizing aluminum.)

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

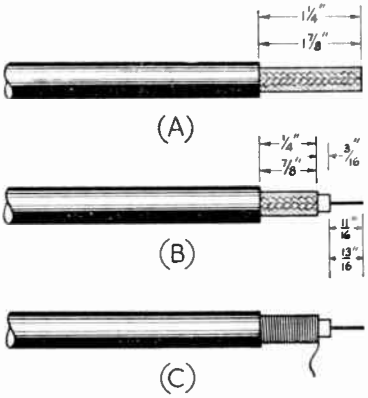


Fig. 24-3 — **Cable-stripping dimensions for Jones Type P-101 plugs. Smaller dimensions are for ¼-inch plugs, the larger dimensions for ½-inch plugs. As indicated in C, the remaining copper braid is wound with bare or tinned wire to make a snug fit in the sleeve of the plug.**

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.

Wiring

If the plate voltage exceeds 500, special care should be used in selecting wire that has adequate voltage rating for transmitter plate-power circuits. To be conservative, the insulation should be good for twice the plate voltage and this figure should be doubled again if the lead carries modulation as well as d.c.

The wire for filament circuits should be of sufficient size to assure rated voltage at the filament terminals (see wire table in the miscellaneous-data chapter).

Power wiring in transmitters should be shielded as a means of reducing TVI. If shielded wire with sufficient voltage rating is not available, unshielded wire can be covered with copper braid.

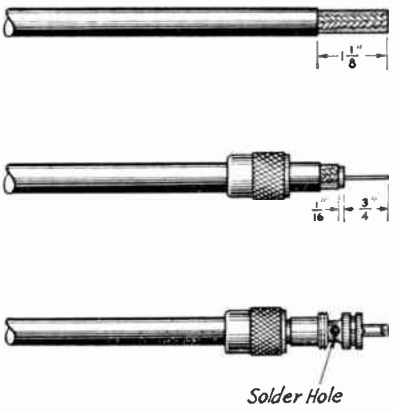


Fig. 24-4 — **Dimensions for stripping ½-inch cable to fit Amphenol Type 83-1SP plug.**

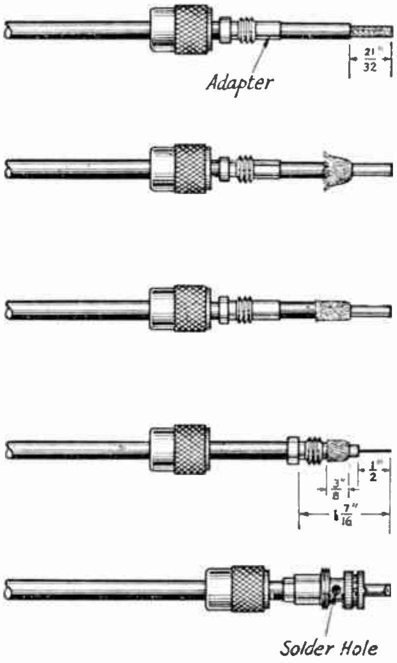


Fig. 24-5 — **Method of assembling ¼-inch cable, Amphenol Type 83-1SP plug and adapter.**

In making connections with shielded wire, the braid should be stripped back about an inch at each end to provide adequate insulation between the conductor and the braid. After fraying the braid back and snipping it off, it should be bound with a few turns of small bare wire, leaving a lead of a few inches for ground to the chassis at the nearest point. Solder should be flowed into the winding, being careful not to apply too much heat that might damage the insulation. The braid of shielded power wires running parallel should be bonded together by spots of solder at frequent intervals. Wires that cross should be similarly bonded. In cases where power-supply leads in the chassis have several branches, it is convenient

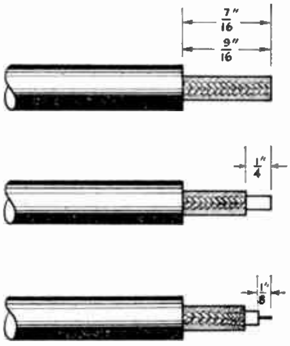


Fig. 24-6 — **Stripping dimensions for Amphenol 82-830 and 82-832 plug-in connectors. The longer exposed braid is for the first type.**

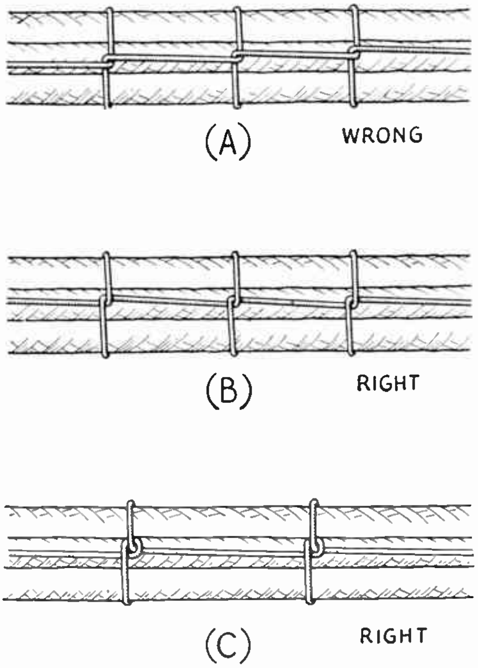


Fig. 24-7 — **Methods of lacing cables. The method shown at C is more secure, but takes more time than the method of B. The latter is usually adequate for most amateur requirements.**

to use fiber terminal or lug strips as anchorages or junction points. Strips of this type are also useful as insulated anchorages for resistors, r.f. chokes and condensers. High-voltage wiring should have exposed points held to a minimum and those which cannot be avoided should be rendered as inaccessible as possible to accidental contact or short-circuit.

For r.f. wiring, soft-drawn bare solid tinned antenna wire, No. 14 or No. 12, is most suitable. Kinks can be removed by stretching a length of 10 or 15 feet and then cutting into short pieces that can be handled conveniently. R.f. wiring should be run directly from point to point over the shortest path and should be kept well spaced from the chassis and components. Where the wiring must pass through the chassis, a half-inch hole should be cut at the appropriate point and the hole lined with a rubber grommet. The wire should not be allowed to touch the rubber.

Power and control wiring outside the transmitter chassis should be cabled to make a neat-looking job. Fig. 24-7 shows the correct methods of lacing cables.

Coaxial Plug Connections

Considerable time and trouble can be saved in making cable connections to coaxial plugs by starting out with the correct stripping dimensions. Fig. 24-3 shows how the end of the cable should be prepared for connecting to Jones Type P-101 plugs. After the exposed braid has been wound, it should be carefully tinned, applying no more heat than is necessary, to avoid melting the inner insulation. A small amount of solder also should be flowed into the sleeve of the plug. Then, when the cable is inserted in the sleeve, the connection can be made secure by holding the iron against the sleeve until the solder inside melts. While joining the two, the plug may be held by inserting it in a hole drilled in a board. Figs. 24-4, 24-5 and 24-6 show details of connections to different types of Amphenol plugs and adapters.

● 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 24-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

TABLE 24-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

would be in the range from 4500 to 4900 ohms, approximately.

Only those values shown in the first column of Table 24-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 24-III.

Fixed Condensers

The methods of marking "postage-stamp" mica condensers, molded paper condensers, and tubular ceramic condensers are shown in Fig. 24-5. 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 ±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 multipler 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.

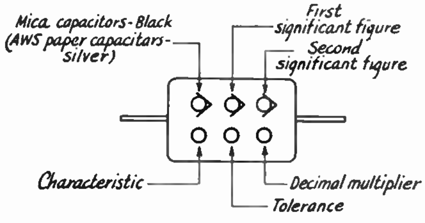
Ceramic Condensers

Conventional markings for ceramic condensers are shown in the lower drawing of Fig. 24-8. The colors have the meanings indicated in Table 24-IV. In practice, dots may be used instead of the *narrow* bands indicated in Fig. 24-8.

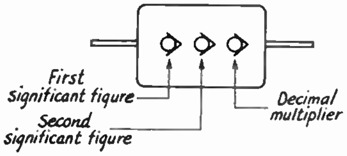
Example: A ceramic condenser has the following markings: Broad band, violet; narrow bands or dots, green, brown, black, green. The significant figures are 5, 1 (51) and the decimal multiplier is 1, so the capacitance is 51 μμfd. The temperature coefficient is − 750 parts per million per degree C., as given by the broad band, and the capacitance tolerance is ± 5%.

Fixed Composition Resistors

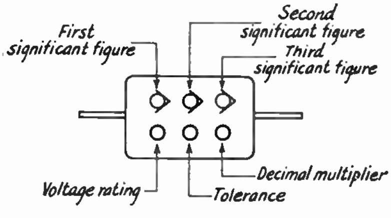
Composition resistors (including small wire-wound units molded in cases identical with the



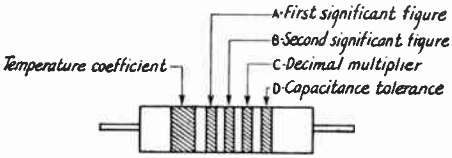
AWS and JAN fixed capacitors



RMA 3-dot 500-volt, ± 20% tolerance only

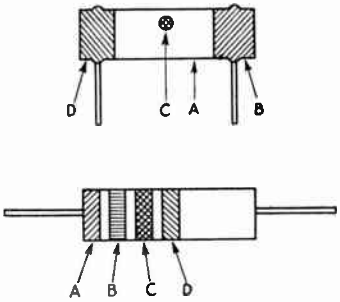


RMA 6-dot



Fixed ceramic capacitors

Fig. 24-8 — Color coding of fixed mica, molded paper, and tubular ceramic condensers. The color code for mica and molded paper condensers is given in Table 24-III. Table 24-IV gives the color code for tubular ceramic condensers.



Fixed composition resistors

Fig. 24-9 — Color coding of fixed composition resistors. The color code is given in Table 24-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%.

composition type) are color-coded as shown in Fig. 24-9. 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. 24-9 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.

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.

TABLE 24-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.

TABLE 24-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

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*

CHAPTER 25

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: W0BY W0BY W0BY W0BY W0BY DE W1AW W1AW [etc.] 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 but does not receive the call letters of the station calling, 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 $\overline{\text{AR}}$, K, $\overline{\text{KN}}$, $\overline{\text{SK}}$ and CL ending signals is as follows:

$\overline{\text{AR}}$ — End of transmission. Recommended after call to a specific station before contact has been established.

> *Example:* W6ABC W6ABC W6ABC DE W9LMN W9LMN W9LMN $\overline{\text{AR}}$. 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.

$\overline{\text{KN}}$ — 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 $\overline{\text{KN}}$.

$\overline{\text{SK}}$ — End of QSO. Recommended before signing *last* transmission at end of a QSO.

Example: .... $\overline{\text{SK}}$ 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: .... $\overline{\text{SK}}$ 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 facilitates break-in operation. It is only necessary with break-in to pause just a moment with the key 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.

With a tap of the key, the man on the receiving end can interrupt (if a word is missed). The other operator is constantly monitoring, awaiting just such directions. 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) press 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-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!

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.

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
BAKER
CHARLIE
DAVID
EDWARD
FRANK
GEORGE
HENRY
IDA
JOHN
KING
LEWIS
MARY
NANCY
OTTO
PETER
QUEEN
ROBERT
SUSAN
THOMAS
UNION
VICTOR
WILLIAM
X-RAY
YOUNG
ZEBRA

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.

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 $\overline{SK}$, 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 $\overline{SK}$ 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 $\overline{KN}$, $\overline{AR}$, 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.

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:

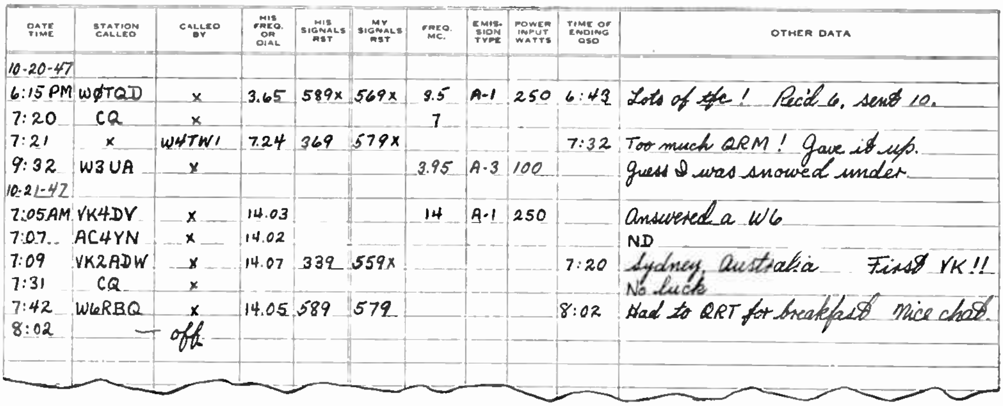
HM — Will start to listen at *high*-frequency end of band and tune toward *middle* of band.
MH — Will start to listen in the *middle* of the band and tune toward the *high*-frequency end.
LM — Will start to listen at *low*-frequency end of band and tune toward *middle* of band.
ML — Will start to listen in the *middle* of the band and 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. This accounts in part for the relative popularity of the 14- and 28-Mc. bands among amateurs who like to work DX.



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	W0TQD	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										
7:05AM	VK4DV	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.

● 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 non-amateur 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.

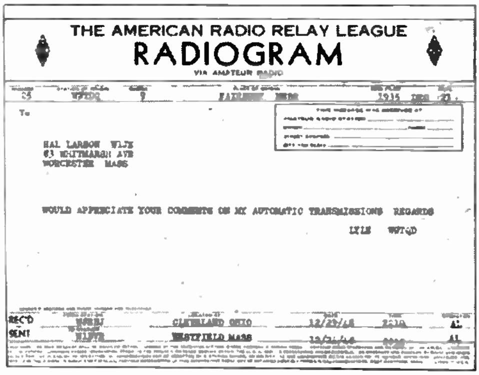
Responsibility

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 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 this Chapter.



THE AMERICAN RADIO RELAY LEAGUE

RADIOGRAM

VIA AMATEUR RADIO

To

HAL LARSON W1JX
63 WHITMARSH AVE
WORCESTER MASS

WOULD APPRECIATE YOUR COMMENTS ON MY AUTOMATIC TRANSMISSIONS REGARDS

LYLE W9TQD

RECD CLEVELAND OHIO

SENT WESTFIELD MASS

Here is an example of a plain-language message in correct ARRL form.

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, there is in existence 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 2130 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. It is open to any amateur who wishes to participate.

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.

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;

EMERGENCY RADIO UNIT

Registered with the Amateur Radio Emergency Corps

Sponsored by the American Radio Relay League in the Public Interest

This is one of the aids which are available through your local EC to put amateur radio before the public eye. It can be used on mobile rigs or self-contained portable stations. Your EC has or can get a supply of them from ARRL headquarters.

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 coordinating 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

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.

local organization are members of the Amateur Radio Emergency Corps (AREC). *All* amateurs are invited to register in the AREC, 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 Coordinator, 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.

For many years amateurs have been serving the public in man-made and natural disasters of various kinds. Now, as once before in our history, we are being called upon to prepare for participation in civil defense communication. The contribution our larger number of mobiles and emergency powered facilities can make is today a much greater one than ever before. Within the limits set by security and frequency availabilities, we hope that we might have an opportunity to serve civil defense needs much more effectively than before. The need is greater than ever. To counterbalance this, our Emergency Corps organization is much stronger, much larger and much more efficient than ever before in its history.

The extent to which we will figure in the completed plans for civil defense at all levels depends entirely on the extent to which we participate as an organization, as one strong facility, in the plans and preparations being made. And while we are doing this we must not forget that it by no means relieves us of the responsibility for continuing to carry on our traditional preparation for and participation in peacetime communications emergencies. We simply have an extra job to do.

Among the League's publications is a booklet entitled *Emergency Communications*. This booklet, while small in size, contains a wealth of information on AREC organization and functions and is invaluable to any amateur participating in emergency work. It is free to AREC 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 this chapter.

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.

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 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.

The American Radio Relay League, Inc.

COMMUNICATIONS DEPARTMENT

Official Relay Station

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, EC's, RMs, PAMs may wear the emblem with green background. Observers and all station appointees are entitled to wear emblems with blue background.

● SECTION NETS

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 into 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.

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 several hundred active 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, 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 year. 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.

The American Radio Relay League, Inc.

OPERATING
ACHIEVEMENT AWARD

This Certifies

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 by the applicant 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 and W0TQD.

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, 13 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.

The American Radio Relay League, Inc.

Certificate of Proficiency in Reception of the Continental Code

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 may 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

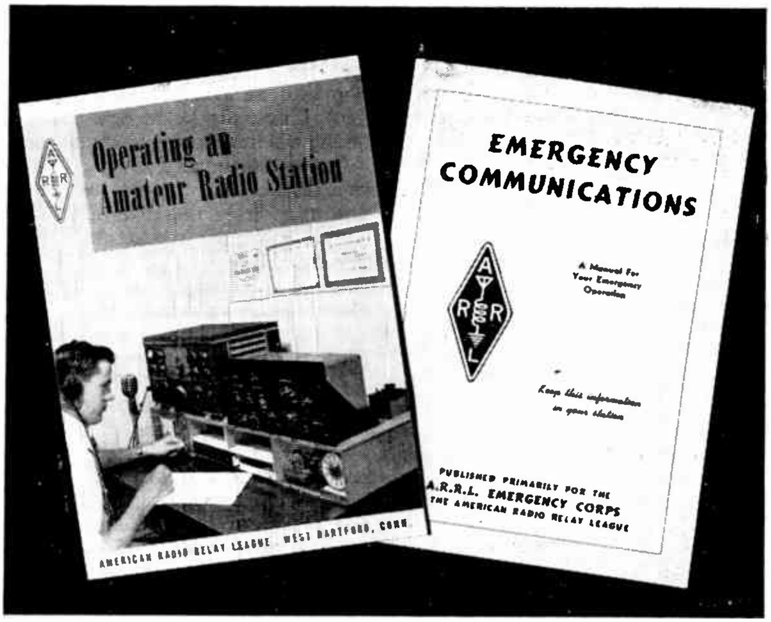
The Old Timers Club is open to anyone who holds an amateur call at the present time, and who held an amateur license (operator or station) 20-or-more years ago. Lapses in activity during the intervening years are permitted.

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 to ARRL Headquarters 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 League 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.

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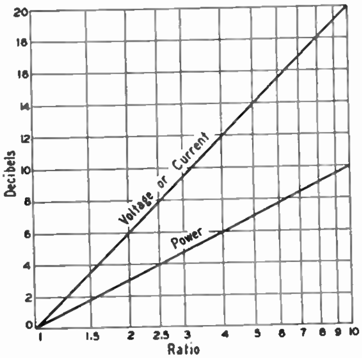
CHAPTER 26

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.

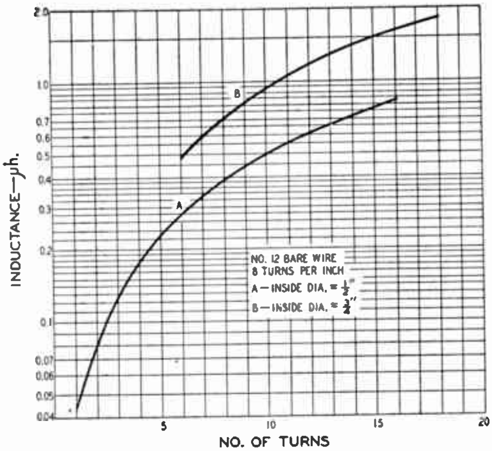
Note that the decibel is based on *power* ratios. Voltage or current ratios can be used, but *only when the impedance is the same for both values of voltage, or current.* The gain of an amplifier cannot be expressed correctly in db. if it is based on the ratio of the output voltage to the input voltage 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.

INDUCTANCE OF SMALL COILS

Most inductance formulas lose accuracy when applied to the small coils used in v.h.f. work and in low-pass filters built for reducing harmonic interference to television, because the conductor thickness is no longer negligible in comparison with the size of the coil. The accompanying chart shows the measured inductance of typical coils used for these purposes, and may be used as a basis for circuit design. Two curves are given: curve A is for coils wound to an inside diameter of ½ inch; curve B is for coils of ¾ inch inside diameter. In both curves the wire size is No. 12, winding pitch 8 turns to the inch (⅛ inch center-to-center turn spacing). The inductance values given include leads ½ inch long.



Measured inductance of coils wound with No. 12 bare wire, 8 turns to the inch. The values include half-inch leads. Where smaller inductance values are required, they should be obtained experimentally by adjusting to the proper resonance frequency with the specified capacitance. Coils of larger inductance can be wound from the common formulas.

SYMBOLS FOR ELECTRICAL QUANTITIES

Admittance	Y, y
Angular velocity (2πf)	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.

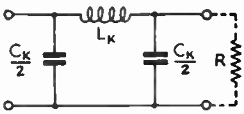
[4] Replace with No. **48**.

[5] Use in 2.5-volt sets where regular bulb burns out too frequently.

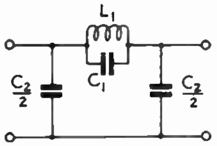
ABBREVIATIONS FOR ELECTRICAL AND RADIO TERMS

Alternating current	a.c.	Medium frequency	m.f.
Ampere (amperes)	a.	Megacycles (per second)	Mc.
Amplitude modulation	AM	Megohm	MΩ
Antenna	ant.	Meter	m.
Audio frequency	a.f.	Microfarad	μfd.
Centimeter	cm.	Microhenry	μh.
Continuous waves	c.w.	Micromicrofarad	μμfd.
Cycles per second	c.p.s.	Microvolt	μv.
Decibel	db.	Microvolt per meter	μv.m.
Direct current	d.c.	Microwatt	μw.
Electromotive force	e.m.f.	Milliampere	ma.
Frequency	f.	Millivolt	mv.
Frequency modulation	FM	Milliwatt	mw.
Ground	gnd.	Modulated continuous waves	m.c.w.
Henry	h.	Ohm	Ω
High frequency	h.f.	Power	P
Intermediate frequency	i.f.	Power factor	p.f.
Interrupted continuous waves	i.c.w.	Radio frequency	r.f.
Kilocycles (per second)	kc.	Ultrahigh frequency	u.h.f.
Kilovolt	kv.	Very-high frequency	v.h.f.
Kilowatt	kw.	Volt (volts)	v.
Magnetomotive force	m.m.f.	Watt (watts)	w.

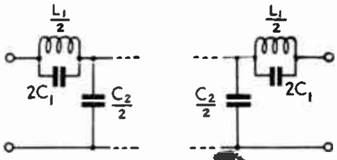
LOW-PASS FILTERS



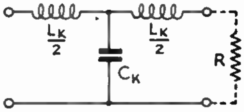
Constant-k π section



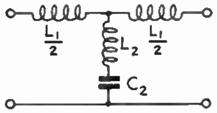
m-derived π section



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

$L_K = \frac{R}{\pi f_c}$ $C_K = \frac{1}{\pi f_c R}$

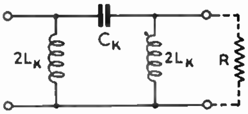
$L_1 = mL_K$ $C_1 = \frac{1-m^2}{4m} C_K$

$L_2 = \frac{1-m^2}{4m} L_K$ $C_2 = m C_K$

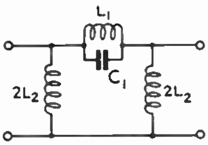
$L_1 = m L_K$ $C_1 = \frac{1-m^2}{4m} C_K$

$L_2 = \frac{1-m^2}{4m} L_K$ $C_2 = m C_K$

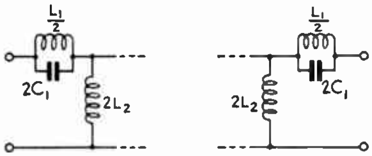
HIGH-PASS FILTERS



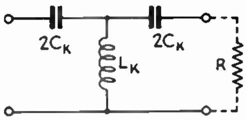
Constant-k π section



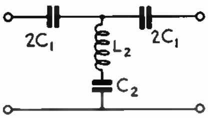
m-derived π section



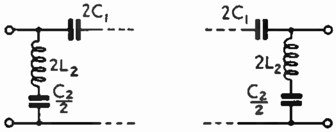
m-derived end sections for use with intermediate π section



Constant-k T section



m-derived T section



m-derived end section for use with intermediate T section

$L_K = \frac{R}{4\pi f_c}$ $C_K = \frac{1}{4\pi f_c R}$

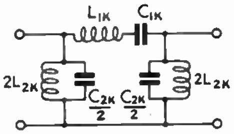
$L_1 = \frac{4m}{1-m^2} L_K$ $C_1 = \frac{C_K}{m}$

$L_2 = \frac{L_K}{m}$ $C_2 = \frac{4m}{1-m^2} C_K$

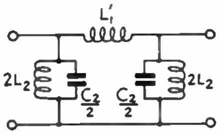
$L_1 = \frac{4m}{1-m^2} L_K$ $C_1 = \frac{C_K}{m}$

$L_2 = \frac{L_K}{m}$ $C_2 = \frac{4m}{1-m^2} C_K$

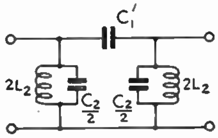
BANDPASS FILTERS



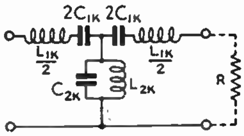
Constant-k π section



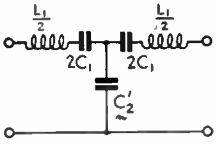
Three-element π section



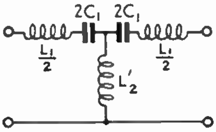
Three-element π section



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

$$m = \sqrt{1 - \left(\frac{f_c}{f_\infty}\right)^2} \text{ for the low-pass filter and}$$

$$m = \sqrt{1 - \left(\frac{f_\infty}{f_c}\right)^2} \text{ for the high-pass filter.}$$

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.

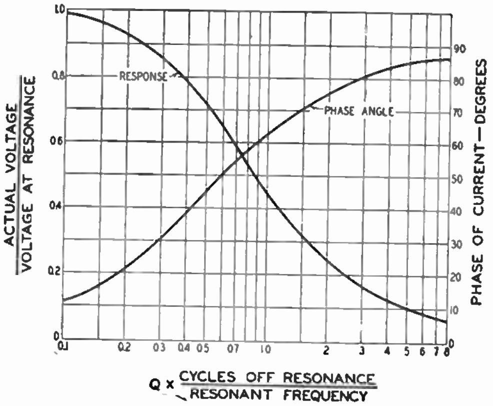
Mann, "An Inexpensive Sideband Filter," *QST*, March, 1949.

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.



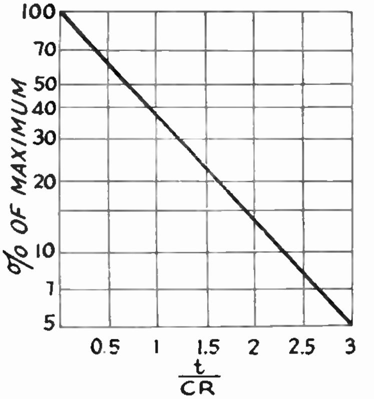
Circuit Q is equal to

$$2\pi fRC \quad \text{or} \quad \frac{R}{2\pi fL}$$

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 μμfd.

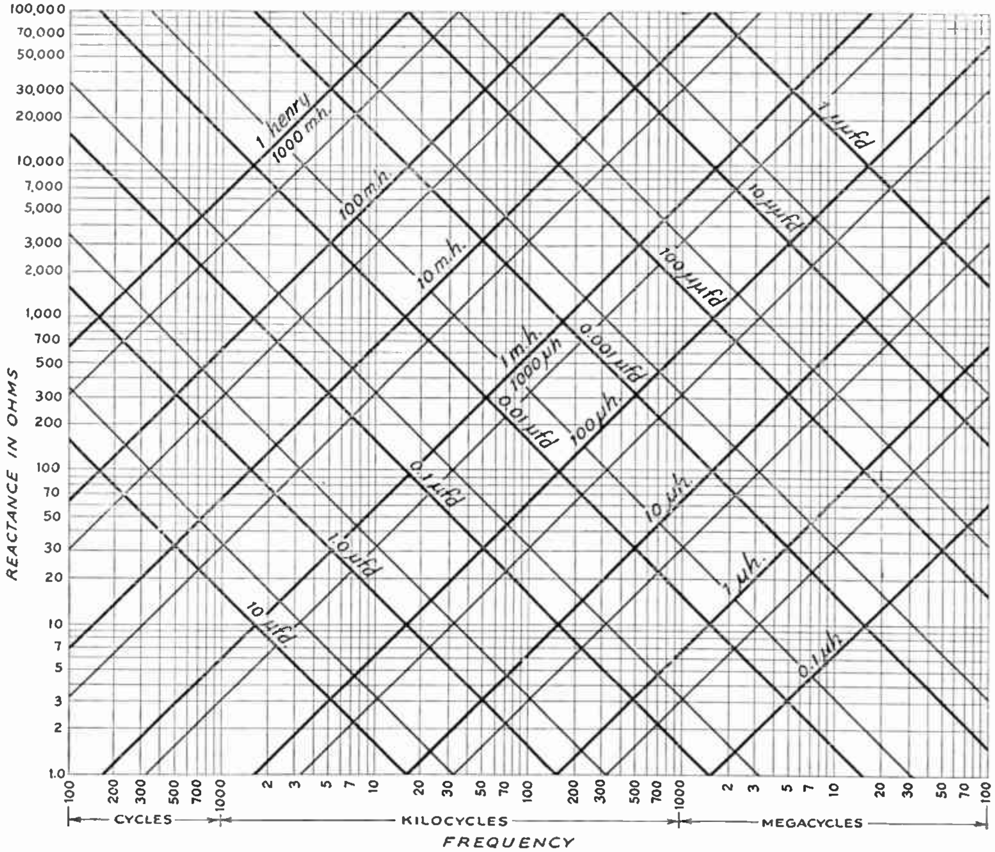
[1] Matched dual diode. [2] Unit has four matched diodes. [3] Manufactured by GE. Other units by Sylvania.

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
Zinx	28.2	0.0035

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.

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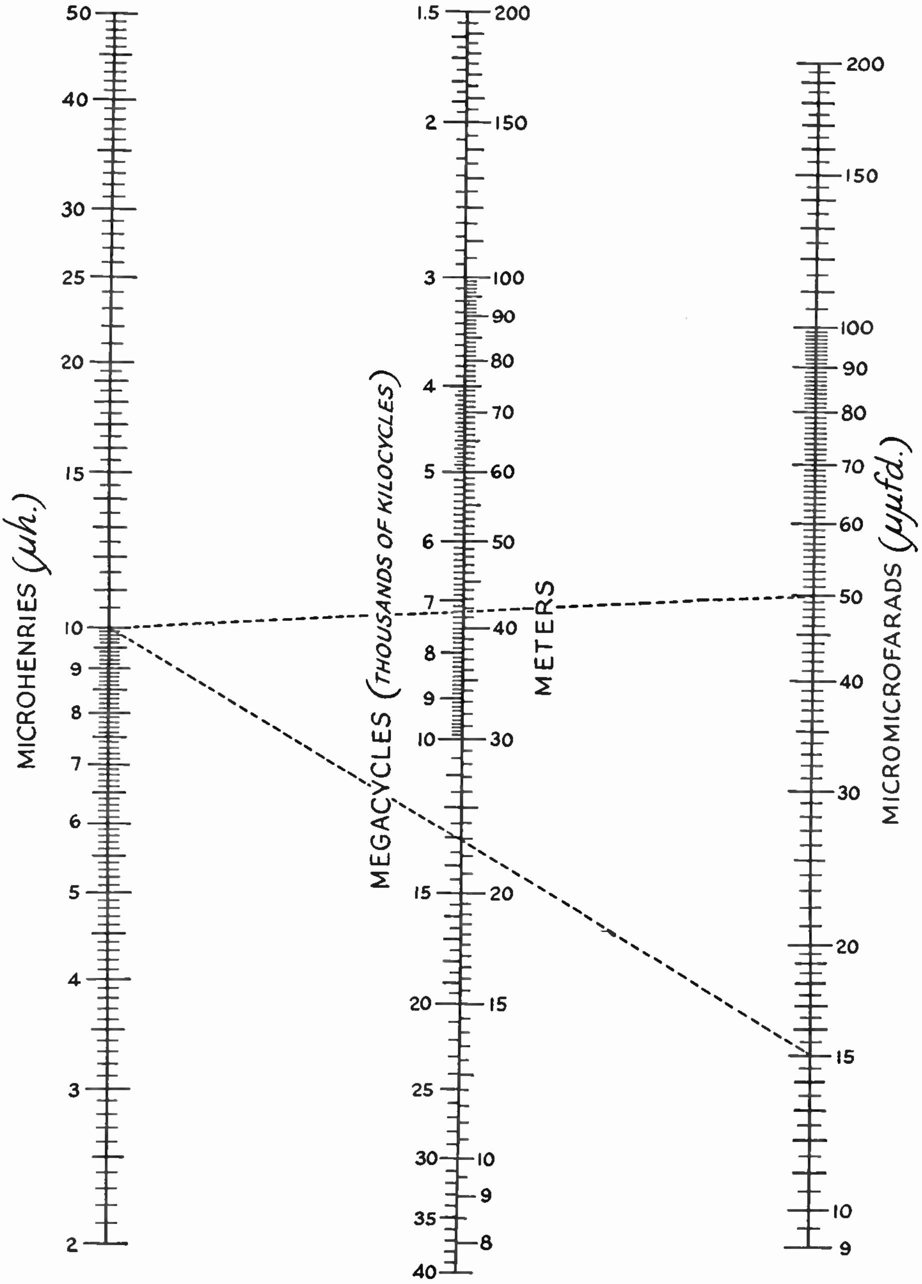
[1] At 20° C., based on copper as 100. [2] Per °C. at 20° C.

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 422 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

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[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.

(Bottom Octave)

	Note	*Frequency*
	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

	Note	*Frequency*
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	W0
Connecticut	W1
Delaware	W3
District of Columbia	W3
Florida	W4
Georgia	W4
Idaho	W7
Illinois	W9
Indiana	W9
Iowa	W0
Kansas	W0
Kentucky	W4
Louisiana	W5
Maine	W1
Maryland	W3
Massachusetts	W1
Michigan	W8
Minnesota	W0
Mississippi	W5
Missouri	W0
Montana	W7
Nebraska	W0
Nevada	W7
New Hampshire	W1
New Jersey	W2
New Mexico	W5
New York	W2
North Carolina	W4
North Dakota	W0
Ohio	W8
Oklahoma	W5
Oregon	W7
Pennsylvania	W3
Rhode Island	W1
South Carolina	W4
South Dakota	W0
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	Ethiopia
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*

AC3	Sikkim
AC4	Tibet
AG2	(See I)
AP	Pakistan
AR8	Lebanon
C (unofficial)	China
C3	Formosa
C9	Manchuria
CE	Chile
CM, CO	Cuba
CN	French Morocco
CP	Bolivia
CR4	Cape Verde Islands
CR5	Portuguese Guinea
CR5	Principe, Sao Thome
CR6	Angola
CR7	Mozambique
CR8	Goa (Portuguese India)
CR9	Macau
CR10	Portuguese Timor
CT1	Portugal
CT2	Azores Islands
CT3	Madeira Islands
CX	Uruguay
DL	Germany
DU	Philippine Islands
EA	Spain
EA6	Balearic Islands
EA8	Canary Islands
EA9	Spanish Morocco
EI	Eire (Irish Free State)
EK	Tangier Zone
EL	Liberia
EP, EQ	Iran (Persia)
ET	Ethiopia
F	France
FA	Algeria
FB8	Kerguelen Islands & New Amsterdam Island
FB8	Madagascar
FC	Corsica
FD8	French Togoland
FE8	French Cameroons
FF8	French West Africa
FG8	Guadeloupe
FI8	French Indo-China
FK8	New Caledonia
FL8	French Somaliland
FM8	Martinique
FN	French India
FO8	French Oceania (e.g., Tahiti)
FP8	St. Pierre & Miquelon Islands
FQ8	French Equatorial Africa
FR8	Reunion Island
FU8, YJ	New Hebrides
FY8	French Guiana & Inini
G	England
GC	Channel Islands
GD	Isle of Man
GI	Northern Ireland
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
I, AG2, MF2	Trieste
I5, MD4, MS4	Italian Somaliland
I6, MD3, MI3	Eritrea
IS	Sardinia
IT1	Sicily
J	Japan
JA0	Bonin & Volcano Islands
K	(See W)
KB6	Baker, Howland & American Phoenix Islands
KC6	Caroline Islands
KC6	Palau Islands
KG4	Guantanamo Bay
KG6	Mariana Islands
KH6	Hawaiian Islands
KJ6	Johnston Island
KL7	Alaska
KM6	Midway Islands
KP4	Puerto Rico
KP6	Palmyra Group, Jarvis Island
KR6	Ryukyu Islands (e.g., Okinawa)
KS4	Swan Island
KS6	American Samoa
KV4	Virgin Islands
KW6	Wake Island
KX6	Marshall Islands
KZ5	Canal Zone
LA	Norway
LA	Svalbard (Spitzbergen)
LI, MC1, 2, MD1, 2, MT1, 2	Libya
LU	Argentina
LX	Luxembourg
LZ	Bulgaria
M1	San Marino
MB9	(See OE)
MC1, 2	(See LI)
MD1, 2	(See LI)
MD3	(See I6)
MD4	(See I5)
MD5	(See SU)
MD6	(See YI)
MD7	(See ZC4)
MF2	(See I)
MI3	(See I6)
MP4	(See VU7)
MP4	Oman
MS4	(See I5)
MT1, 2	(See LI)
OA	Peru
OE, MB9	Austria
OH	Finland
OK	Czechoslovakia
ON	Belgium
OQ	Belgian Congo
OX	Greenland
OY	Faeroes
OZ	Denmark
PA	Netherlands
PJ	Netherlands West Indies
PK1, 2, 3	Java
PK4	Sumatra
PK5	Netherlands, Borneo
PK6	Celebes & Molucca Islands
PK6	Netherlands New Guinea
PX	Andorra
PY	Brazil
PZ	Netherlands Guiana
SM	Sweden
SP	Poland
ST	Anglo-Egyptian Sudan
SU, MD5	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, 0	Asiatic Russian S.F.S.R.
UB5	Ukraine
UC2	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
UP2	Lithuania
UQ2	Latvia
UR2	Estonia
VE, VO	Canada
VK	Australia (including Tasmania)
VK1	Heard Island
VK1	Macquarie Island
VK9	Papua Territory
VK9	Territory of New Guinea
VK9	Norfolk Island
VO	(See VE)
VP1	British Honduras
VP2	Leeward Islands
VP2	Windward Islands
VP3	British Guiana
VP4	Trinidad & Tobago
VP5	Cayman Islands
VP5	Jamaica
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	British Somaliland
VQ8	Chagos Islands
VQ8	Mauritius
VQ9	Seychelles
VR1	Gilbert & Ellice Islands & Ocean Island
VR1	British Phoenix Islands
VR2	Fiji Islands
VR3	Fanning Island (Christmas Island)
VR4	Solomon Islands
VR5	Tonga (Friendly) Islands
VR6	Pitcairn Island
VS1, 2	Malaya
VS4	British North Borneo
VS5	Brunei
VS5	Sarawak
VS6	Hong Kong
VS7	Ceylon
VS9	Aden & Socotra
VS9	Maldive Islands
VU	India
VU4	Laccadive Islands
VU7, MP4	Bahrein Islands
W, K	United States of America
XE	Mexico
XZ	Burma
YA	Afghanistan
YI, MD6	Iraq
YJ	(See FU8)
YK	Syria
YN	Nicaragua
YO, YR	Roumania
YS	Salvador
YT, YU	Yugoslavia
YV	Venezuela
ZA	Albania
ZB1	Malta
ZB2	Gibraltar
ZC1	Transjordan
ZC2	Cocos Islands
ZC3	Christmas Island
ZC4, MD7	Cyprus
ZC6	Palestine
ZD1	Sierra Leone
ZD2	Nigeria
ZD3	Gambia
ZD4	Gold Coast, Togoland
ZD6	Nyasaland
ZD7	St. Helena
ZD8	Ascension Island
ZD9	Tristan da Cunha & Gough Island
ZE	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
3A1, 2	Monaco
3V8	Tunisia
4X4	Israel
9S4	Saar
	Aldabra Islands
	Andaman and Nicobar Islands
	Antarctica
	Bhutan
	Clipperton Island
	Comoro Islands
	Easter Island
	Fridtjof Nansen Land (Franz Josef Land)
	Galapagos Islands
	Ifni
	Jan Mayen Island
	Kuwait
	Marion Island
	Mongolia
	Nepal
	Rio de Oro
	Spanish Guinea
	Tannu Tuva
	Tokelau (Union) Islands
	Wrangel Islands
	Yemen

CHAPTER 27

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 V13–V57) and the identifying base-diagram number in the base-diagram section (pages V5–V12).

Type	*Page*	*Base*
00-A	V24	4D
01-A	V24	4D
0A2	V31	5BO
0A3	V32	4AJ
0A4G	V31	4V
0A5	V31	Fig. 33
0B2	V31	5BO
0B3	V32	4AJ
0C3	V32	4AJ
0D3	V32	4AJ
0Y4	V36	4BU
0Z4	V36	4R
0Z4A	V37	4R
1	V36	4G
1A3	V26	5AP
1A4P	V19	4M
1A4T	V19	4K
1A5GT	V21	6X
1A6	V19	6L
1A7GT	V21	7Z
1AB5	V21	5BF
1AC5	V29	Fig. 14
1AD5	V29	Fig. 16
1AF4	V26	6AR
1AF5	V26	6AU
1B3GT	V36	3C
1B4P	V19	4M
1B5	V19	6M
1B7GT	V21	7Z
1B8GT	V21	8AW
1B47	V31	—
1B48	V36	—
1C3	V26	5CF
1C5GT	V21	6X
1C6	V20	6L
1C7G	V20	7Z
1C8	V29	—
1C21	V31	4V
1D5GP	V20	5Y
1D5GT	V20	5R
1D7G	V20	7Z
1D8GT	V21	8AJ
1E4G	V21	5S
1E5GP	V20	5Y
1E7G	V20	8C
1E8	V29	Fig. 27
1F4	V20	5K
1F5G	V20	6X
1F6	V20	6W
1F7G	V20	7AD
1G4GT	V21	5S
1G5G	V20	6X
1G6GT	V21	7AB
1H4G	V20	5S
1H5GT	V21	5Z
1H6G	V20	7AA
1J5G	V20	6X
1J6G	V20	7AB
1L4	V26	6AR
1L6	V26	7DC
1LA4	V21	5AD
1LA6	V21	7AK
1LB4	V21	5AD
1LB6	V21	8AX
1LC5	V21	7AO
1LC6	V21	7AK
1LD5	V21	6AX
1LE3	V21	4AA
1LG5	V21	7AO
1LH4	V21	5AG
1LN5	V21	7AO
1N5GT	V21	5Y
1N6G	V21	7AM
1P5GT	V21	5Y
1Q5GT	V21	6AF
1R4	V21	4AH
1R5	V26	7AT
1S4	V26	7AV
1S5	V26	6AU
1SA6GT	V21	6CA
1SB6GT	V21	6CB
1T4	V26	6AR
1T5GT	V21	6AF
1T6	V29	Fig. 28
1U4	V26	6AR
1U5	V26	6BW
1U6	V26	7DC
1-V	V36	4G
1V2	V36	9C
1V5	V29	—
1W4	V26	5BZ
1W5	V29	—
1X2	V36	9Y
1X2A	V36	9Y
1Z2	V36	7CB
2A3	V19	4D
2A4G	V31	5S
2A5	V19	6B
2A6	V19	6G
2A7	V19	7C
2AP1	V32	11B
2B4	V32	5A
2B6	V19	7J
2B7	V19	7D
2B22	V14	Fig. 37
2B25	V36	3T
2BP1	V32	12E
2C4	V32	5AS
2C21	V17	7BH
2C21	V39	7BH
2C22	V14	4AM
2C22	V39	4AM
2C25	V40	4D
2C26A	V39	4BB
2C34	V39	T-7DC
2C35	V25	Fig. 38
2C36	V39	Fig. 36
2C37	V39	Fig. 36
2C39	V46	—
2C39A	V46	—
2C40	V39	Fig. 19
2C43	V39	Fig. 19
2C44	V25	Fig. 17
2C51	V26	8CJ
2D21	V32	7BN
2E5	V19	6R
2E22	V52	5J
2E24	V50	7CL
2E25	V51	5BJ
2E26	V50	7CK
2E30	V26	7CQ
2E30	V50	7CQ
2E31	V29	—
2E32	V29	—
2E35	V29	—
2E36	V29	—
2E41	V29	—
2E42	V29	—
2G5	V19	6R
2G21	V29	—
2G22	V29	—
2J42	V57	—
2J42A	V57	—
2K25	V55	Fig. 60
2K26	V55	Fig. 60
2K28	V55	Fig. 61
2K33	V55	Fig. 62
2K34	V55	Fig. 58
2K35	V55	Fig. 58
2K39	V55	Fig. 59
2K41	V55	Fig. 59
2K42	V55	Fig. 59
2K43	V55	Fig. 59
2K44	V55	Fig. 59
2K46	V55	Fig. 58
2K47	V55	Fig. 58
2K56	V55	Fig. 60
2S/4S	V19	5D
2V3G	V36	4Y
2W3	V36	4X
2X2	V36	4AB
2X2-A	V36	4AB
2Y2	V36	4AB
2Z2	V36	4B
3A4	V26	7BB
3A4	V50	7BB
3A5	V26	7BC
3A5	V39	7BC
3A8GT	V24	8AS
3AP1	V33	7AN
3B4	V50	7CY
3B5GT	V24	7AP
3B7	V21	7BE
3B7	V39	7BE
3B24	V36	T-4A
3B25	V36	4P
3B26	V36	Fig. 31
3B27	V36	4P
3BP1	V33	14A
3C5GT	V24	7AQ
3C6	V24	7BW
3C22	V16	Fig. 30
3C23	V32	3G
3C24	V41	2D
3C28	V41	Fig. 56
3C34	V41	3G
3C37	V47	—
3D6	V21	6BB
3D6	V50	6BB
3D23	V52	Fig. 54
3D24	V53	T-9J
3DP1	V33	Fig. 49
3DX3	V52	Fig. 40
3E5	V26	6BX
3E6	V21	7CJ
3E22	V52	8BY
3E29	V52	7BP
3EP1	V33	11A
3GP1	V33	11A
3J31	V57	—
3JP1	V33	14B
3K21	V55	Fig. 58
3K22	V55	Fig. 58
3K23	V55	Fig. 59
3K27	V56	Fig. 59
3K30	V56	Fig. 58
3KP1	V33	11M
3LE4	V24	6BA
3LF4	V24	6BB
3MP1	V33	Fig. 2
3Q4	V26	7BA
3Q5GT	V24	7AQ
3RP1	V33	12E
3S4	V26	7BA
3V4	V26	6BX
3-25A3	V40	3G
3-25D3	V41	2D
3-50A4	V42	3G
3-50D4	V42	2D
3-50G2	V42	2D
3-75A2	V44	2D
3-75A3	V44	2D
3-100A2	V45	2D
3-100A4	V45	2D
3X-100A11	V46	—
3-150A2	V47	4BC
3-150A3	V47	4BC
3X-150A3	V47	—
3-200A3	V48	Fig. 52
3-250A2	V48	2N
3-250A4	V48	2N
3-300A2	V49	4BC
3-300A3	V49	4BC
4A6G	V20	8L
4A6G	V24	8L
4C32	V48	2N
4C34	V48	2N
4C36	V46	Fig. 56
4D21	V54	5BK
4D22	V53	Fig. 50
4D23	V54	5BK
4D32	V53	Fig. 51
4E27	V53	7BM
4E27A	V54	7BM
4J50	V57	—
4J52	V57	—
4J78	V57	—
4X150A	V54	T-9J
4X150G	V54	—
4-65A	V53	5BK
4-125A	V54	5BK
4-250A	V55	5BK
4-400A	V55	5BK
5A6	V50	9L
5AP1	V33	11A
5AZ4	V36	5T
5BP1	V33	11A
5CP1	V33	14B
5C24	V48	Fig. 26
5D22	V55	5BK
5FP1	V33	5AN
5HP1	V33	11A
5JP1	V33	11E
5LP1	V33	11F
5MP1	V33	7AN
5R4GY	V36	5T
5RP1	V33	14F
5T4	V36	5T
5T4P4	V33	12C
5U4G	V36	5T
5UP1	V33	12E
5V4G	V36	5L
5W4	V36	5T
5WP11	V33	12C
5WP15	V34	12C
5X3	V36	4C
5X4G	V36	5Q
5Y3G	V36	5T
5Y4G	V36	5Q
5Z3	V36	4C
5Z4	V36	5L
5-125B	V54	7BM
6A3	V17	4D
6A4	V17	5B
6A5G	V14	6T
6A6	V17	7B
6A7	V17	7C
6A8	V13	8A
6AB4	V26	5CE
6AB5	V17	6R
6AB6G	V14	7AU
6AB7	V13	8N
6AC5GT	V14	6Q
6AC6G	V14	7AU
6AC7	V13	8N
6AD5G	V14	6Q
6AD6G	V14	7AG
6AD7G	V14	8AY
6AE5G	V14	6Q
6AE6GT	V15	7AH
6AE7GT	V15	7AX
6AF5G	V15	6Q
6AF6G	V17	7AG
6AF7G	V15	8AG
6AG5	V26	7BD
6AG6G	V15	7S
6AG7	V13	8Y
6AG7	V50	8Y
6AH5G	V15	6AP
6AH6	V26	7CC
6AH7GT	V15	8BE
6AJ5	V26	7PM
6AJ7	V13	8N
6AK5	V26	7BD
6AK6	V26	7BK
6AK6	V50	7BK
6AK7	V13	8Y
6AL5	V26	6BT
6AL6G	V15	6AM
6AL7GT	V15	8CH
6AM5	V26	6CH
6AM6	V26	7DB
6AN5	V26	7BD
6AN6	V27	7BJ
6AN7	V27	9Q
6AQ5	V27	7BZ
6AQ5	V50	7BZ
6AQ6	V27	7BT
6AQ7GT	V15	8CK
6AR5	V27	6CC
6AR6	V15	6BQ
6AS5	V27	7CV
6AS6	V27	7CM
6AS7G	V15	8BD
6AT6	V27	7BT
6AU5GT	V15	6CK
6AU6	V27	7BK
6AV5GT	V15	6CK
6AV6	V27	7BT
6AX5GT	V36	6S
6AX6G	V36	7Q
6B4G	V15	5S
6B5	V18	6AS
6B6G	V15	7V
6B7	V18	7D
6B8	V13	8E
6BA6	V27	7CC
6BA7	V27	8CT
6BC5	V27	7BD
6BD5GT	V15	6CK
6BD6	V27	7CC
6BE6	V27	7CH
6BF5	V27	7BZ
6BF6	V27	7BT
6BG6	V15	5BT
6BH6	V27	7CM
6BJ6	V27	7CM
6BK6	V27	7BT
6BL6	V56	—
6BM6	V56	—
6BN6	V27	7DF
6BN7	V27	Fig. 41
6BT6	V27	7BT
6BQ6GT	V15	6AM
6C4	V39	6BG
6C4	V27	6BG
6C5	V13	6Q
6C6	V18	6F
6C7	V18	7G
6C8G	V15	8G
6CB6	V27	7CM
6CD6G	V15	5BT
6D4	V32	5AY
6D6	V18	6F
6D7	V18	7H
6D8G	V15	8A
6E5	V18	6R
6E6	V18	7B
6E7	V18	7H
6E8G	V15	8O
6EU6	V27	7BT
6F4	V24	7BR
6F4	V39	7BR
6F5	V13	5M
6F6	V13	7S
6F6	V50	7S
6F7	V18	7E
6F8G	V15	8G
6G5	V18	6R
6G6G	V15	7S
6H4GT	V15	5AF
6H5	V18	6R
6H6	V13	7Q
6H8G	V15	8E
6J4	V27	7BQ
6J5	V13	6Q
6J6	V27	7BF
6J6	V39	7BF
6J7	V13	7R
6J8G	V15	8H
6K4	V29	—
6K5GT	V15	5U
6K6GT	V15	7S
6K7	V13	7R
6K8	V13	8K
6L4	V24	7BR
6L5G	V15	6Q
6L6	V13	7AC
6L6	V51	7AC
6L6GX	V51	7AC
6L7	V13	7T
6M5	V27	9M
6M6G	V15	7S
6M7G	V15	7R
6M8GT	V15	8AU
6N4	V27	7CA
6N4	V39	7CA
6N5	V17	6R
6N6G	V15	7AU
6N7	V13	8B
6N7	V39	8B
6N8	V27	9T
6P5GT	V15	6Q
6P7G	V15	7U
6P8G	V15	8K
6Q4	V27	9S
6Q5G	V32	6Q
6Q6G	V15	6Y
6Q7	V13	7V
6R4	V27	9R
6R6G	V15	6AW
6R7	V13	7V
6R8	V27	9E
6S4	V27	9AC
6S6GT	V15	5AK
6S7	V13	7R
6S8GT	V15	8CB
6SA7	V13	8R
6SB7Y	V13	8R
6SC7	V13	8S
6SD7GT	V15	8M
6SE7GT	V15	8N
6SF5	V14	6AB
6SF7	V14	7AZ
6SG7	V14	8BK
6SH7	V14	8BK
6SH7L	V16	8BK
6SJ7	V14	8N
6SJ7Y	V14	8N
6SK7	V14	8N
6SL7GT	V16	8BD
6SN7GT	V16	8BD
6SN7GTA	V16	8BD
6SQ7	V14	8Q
6SR7	V14	8Q
6SS7	V14	8N
6ST7	V14	8Q
6SU7GTY	V16	8BD
6SV7	V14	7AZ
6SZ7	V14	8Q
6T5	V18	6R
6T6GM	V16	6Z
6T7	V14	7V
6T8	V27	9E
6U4GT	V36	4CG
6U5	V18	6R
6U6GT	V16	7AC
6U7G	V16	7R
6V4	V36	9M
6V6	V14	7AC
6V6GT	V50	7AC
6V7G	V16	7V
6W4GT	V36	4CG
6W5G	V36	6S
6W6GT	V16	7AC
6W7G	V16	7R
6X4	V36	7CF
6X5	V36	6S
6X6G	V16	7AL
6Y3G	V36	4AC
6Y5	V36	6J
6Y6G	V16	7AC
6Y7G	V16	8B
6Z3	V36	4G
6Z4	V37	5D
6Z5	V36	6K
6Z7G	V16	8B
6ZY5G	V36	6S
7A4	V16	5AC
7A5	V16	6AA
7A6	V16	7AJ
7A7	V16	8V
7A8	V16	8U
7AB7	V25	8BO
7AD7	V16	8V
7AF7	V16	8AC
7AG7	V16	8V
7AH7	V16	8V
7AJ7	V16	8V
7AP4	V34	5AJ
7B4	V16	5AC
7B5	V16	6AE
7B6	V16	8W
7B7	V16	8V
7B8	V16	8X
7BP1	V34	5AN
7C4	V25	4AH
7C5	V17	6AA
7C6	V17	8W
7C7	V17	8V
7CP1	V34	6AZ
7D7	V17	8AR
7DP4	V34	12C
7E5	V25	8BN
7E6	V17	8W
7E7	V17	8AE
7EP4	V34	11N
7F7	V17	8AC
7F8	V17	8BW
7G7	V17	8V
7G8	V17	8BV
7GP4	V34	Fig. 47
7H7	V17	8V
7J7	V17	8AR
7JP4	V34	14G

Type	Page	Base
7K7	V17	8BF
7L7	V17	8V
7MP7	V34	12D
7N7	V17	8AC
7Q7	V17	8AL
7R7	V17	8AE
7S7	V17	8BL
7T7	V17	8V
7V7	V17	8V
7W7	V17	8BJ
7X7	V17	8BZ
7Y4	V36	5AB
7Z4	V36	5AB
8AP4	V34	12H
8BP4	V34	14G
9AP4	V34	6AL
9CP4	V34	4AF
9JP1	V34	8BR
10	V24	4D
10	V40	4D
10BP4	V34	12D
10EP4	V34	12D
10FP4	V34	12D
10HP4	V34	14G
10KP7	V34	12D
10Y	V40	4D
11/12	V24	4F
12A4	V27	9AG
12A5	V22	7F
12A6	V22	7AC
12A7	V22	7K
12A7	V36	7K
12A8GT	V22	8A
12AH7GT	V22	8BE
12AL5	V27	6BT
12AP4	V34	6AL
12AT6	V28	7BT
12AT7	V28	9A
12AU6	V28	7CC
12AU7	V28	9A
12AU7	V39	9A
12AV6	V28	7BT
12AW6	V28	7CM
12AW7	V28	7CM
12AX7	V28	9A
12AY7	V28	9A
12B6M	V22	6Y
12B7	V22	8V
12B7ML	V22	8V
12B8GT	V22	8T
12BA6	V28	7CC
12BA7	V28	8CT
12BD6	V28	7CC
12BE6	V28	7CH
12BF6	V28	7BT
12BH7	V28	9A
12BK6	V28	7BT
12BN6	V28	7DF
12BT6	V28	7BT
12C8	V22	8E
12CP4	V34	4AF
12DP4	V34	5AN
12E5GT	V22	6Q
12EU6	V28	7BT
12F5GT	V22	5M
12G7G	V22	7V
12H6	V22	7Q
12J5GT	V22	6Q
12J7GT	V22	7R
12K7GT	V22	7R
12K8	V22	8K
12KP4A	V34	Fig. 35
12L8GT	V22	8BU
12LP4	V34	12D
12Q7GT	V22	7V
12S8GT	V22	8CB
12SA7	V22	8R
12SC7	V22	8S
12SF5	V22	6AB
12SF7	V22	7AZ
12SG7	V22	8BK
12SH7	V22	8BK
12SJ7	V22	8N
12SK7	V22	8N
12SL7GT	V22	8BD
12SN7GT	V22	8BD
12SQ7	V22	8Q
12SR7	V22	8Q
12SW7	V22	8Q
12SX7	V22	8BD
12SY7	V22	8R
12Z3	V36	4G
12Z5	V36	7L
14A4	V22	5AC
14A5	V22	6AA
14A7	V22	8V
14AF7	V22	8AC
14AF7	V24	8AC
14B6	V22	8W
14B8	V22	8X
14BP4	V34	Fig. 35
14C5	V22	6AA
14C7	V23	8V
14E6	V23	8W
14E7	V23	8AE
14F7	V23	8AC
14F8	V23	8BW
14H7	V23	8V
14J7	V23	8BL
14N7	V23	8AC
14Q7	V23	8AL
14R7	V23	8AE
14S7	V23	8BL
14V7	V23	8V
14W7	V23	8BJ
14Y4	V36	5AB
14Z3	V36	4G
15	V20	5F
15AP4	V34	12D
15CP4	V34	Fig. 35
15DP4	V34	12D
15E	V40	T-4AF
16AP4	V34	Fig. 35
16EP4A	V34	12D
16FP4	V34	Fig. 35
16GP4	V34	12D
16HP4	V34	Fig. 35
16JP4	V34	12D
16KP4	V34	12D
16LP4	V34	Fig. 35
16RP4	V34	12D
16TP4	V34	Fig. 35
16UP4	V34	12D
17	V32	3G
18	V23	6B
19	V20	6C
19AP4	V34	Fig. 35
19AP4A	V35	12D
19BG6G	V23	5BT
19FP4	V35	12D
19J6	V28	7BF
19T8	V28	9E
20BP4	V35	12D
20	V24	4D
20J8GM	V23	8H
21A7	V23	8AR
22	V24	4K
22AP4	V35	Fig. 35
24-A	V19	5E
24-G	V41	2D
24XH	V35	Fig. 1
25A6	V23	7S
25A7GT	V36	8F
25A7GT	V23	8F
25AC5G	V23	6Q
25AV5GT	V23	6CK
25B5	V23	6D
25B6G	V23	7S
25B8GT	V23	8T
25BQ6GT	V23	6AM
25C6G	V23	7AC
25D8GT	V23	8AF
25L6	V23	7AC
25N6G	V23	7W
25S	V19	6M
25T	V40	3G
25W4	V36	4CG
25X6GT	V36	7Q
25Y4GT	V36	5AA
25Y5	V36	6E
25Z3	V36	4G
25Z4	V36	5AA
25Z5	V36	6E
25Z6	V36	7Q
26	V24	4D
26A6	V28	7BK
26BK6	V28	7BT
26A7GT	V23	8BU
26C6	V28	7BT
26D6	V28	7CH
27	V19	5A
28D7	V24	8BS
28Z5	V37	5AB
30	V20	4D
31	V20	4D
32	V20	4K
32L7GT	V23	8Z
32L7GT	V37	8Z
33	V20	5K
34	V20	4M
35/51	V19	5E
35A5	V23	6AA
35B5	V28	7BZ
35C5	V28	7CV
35L6GT	V23	7AC
35T	V42	3G
35TG	V42	2D
35W4	V37	5BQ
35Y4	V37	5AL
35Z3	V37	4Z
35Z4GT	V37	5AA
35Z5G	V37	6AD
35Z6G	V37	7Q
36	V18	5E
37	V18	5A
38	V18	5F
39/44	V18	5F
40	V24	4D
40Z5GT	V37	6AD
41	V18	6B
42	V18	6B
43	V23	6B
45	V19	4D
45Z3	V37	5AM
45Z5GT	V37	6AD
46	V19	5C
47	V19	5B
48	V23	6A
49	V20	5C
50	V24	4D
50A5	V23	6AA
50AX6G	V37	7Q
50B5	V28	7BZ
50C5	V28	7CV
50C6GT	V23	7AC
50L6GT	V23	7AC
50T	V44	2D
50X6	V37	7AJ
50Y6GT	V37	7Q
50Y7GT	V37	8AN
50Z6G	V37	7Q
50Z7G	V37	8AN
51	V19	5E
52	V18	5C
53	V19	7B
53A	V41	T-4B
55	V19	6G
56	V19	5A
56AS	V18	5A
57	V19	6F
57AS	V18	6F
58	V19	6F
58AS	V18	6F
59	V19	7A
70A7GT	V23	8AB
70A7GT	V37	8AB
70L7GT	V23	8AA
70L7GT	V37	8AA
71-A	V25	4D
72	V37	4P
73	V37	4Y
75	V18	6G
75TH	V44	2D
75TL	V44	2D
76	V18	5A
77	V18	6F
78	V18	6F
79	V18	6H
80	V37	4C
81	V37	4B
82	V37	4C
83	V37	4C
83-V	V37	4AD
84/6Z4	V37	5D
85	V18	6G
85AS	V18	6G
89	V18	6F
99	V25	4D
100TH	V45	2D
100TL	V45	2D
111H	V44	2D
112-A	V25	4D
117L7GT	V23	8AO
117L7GT	V37	8AO
117M7GT	V23	8AO
117M7GT	V37	8AO
117N7GT	V23	8AV
117N7GT	V37	8AV
117P7GT	V23	8AV
117P7GT	V37	8AV
117Z3	V37	4BR
117Z4GT	V37	5AA
117Z6GT	V37	7Q
128AS	V32	5A
150T	V47	2N
152TH	V47	4BC
152TL	V47	4BC
182-B	V25	4D
183	V25	4D
203-A	V45	4E
203-H	V45	3N
204-A	V49	T-1A
205D	V40	4D
211	V45	4E
212-E	V49	T-2A
217A	V37	4AT
217C	V37	4AT
227A	V45	T-4B
241B	V49	T-2AA
242A	V44	4E
242B	V45	4E
242C	V46	4E
249B	V37	Fig. 53
250TH	V48	2N
250TL	V48	2N
254A	V51	T-4C
254B	V52	T-4C
261A	V46	4E
270A	V49	T-1A
276A	V46	4E
282A	V53	T-4C
284B	V46	3N
284D	V44	4E
295A	V46	4E
300T	V49	2N
303A	V45	4E
304A	V49	T-1A
304B	V42	2D
304TH	V49	4BC
304TL	V49	4BC
305A	V53	T-4CE
306A	V51	T-5CB
307A	V51	T-5C
308B	V49	T-2A
310	V40	4D
311	V45	4E
311CH	V46	Fig. 57
312A	V53	T-6C
312E	V49	T-2AA
316A	V41	—
327A	V45	T-4AD
327B	V44	T-4AD
342B	V45	4E
356A	V43	T-4BD
361A	V46	4E
376A	V46	4E
410R	V56	Fig. 58
482B	V25	4D
483	V25	4D
485	V25	5A
527	V49	T-4B
559	V25	Fig. 18
592	V48	Fig. 52
703A	V40	—
705A	V37	T-3AA
707B	V56	Fig. 61
715B	V53	—
717A	V16	8BK
723AB	V55	Fig. 60
756	V42	4D
800	V41	2D
801	V40	4D
801A	V40	4D
802	V50	6BM
803	V54	5J
804	V53	T-5C
805	V47	3N
806	V48	2N
807	V52	5AW
807W	V52	5AW
808	V43	2D
809	V41	3G
810	V47	2N
811	V43	3G
811A	V43	3G
812	V43	3G
812A	V43	3G
812H	V44	3G
813	V54	5BA
814	V53	T-5D
815	V52	8BY
816	V37	4P
822	V48	3N
822S	V48	2N
826	V43	7BO
828	V54	5J
829	V52	7BP
829A	V52	7BP
829B	V52	7BP
830	V42	4D
830B	V43	3G
831	V49	T-1AA
832	V51	7BP
832A	V51	7BP
833A	V49	T-1AB
834	V42	2D
835	V45	4E
836	V37	4P
837	V50	6BM
838	V46	4E
840	V20	5J
841	V40	4D
841A	V43	3G
841SW	V43	3G
843	V40	5A
844	V51	5AW
849	V49	T-1A
850	V54	T-3B
852	V46	2D
860	V54	T-4CB
861	V55	T-1B
864	V25	4D
865	V51	T-4C
866	V37	4P
866A	V37	4P
866B	V37	4P
866Jr.	V37	4B
871	V37	4P
872	V37	4AT
872A	V37	4AT
874	V32	4S
876	V32	—
878	V37	4P
879	V37	4P
884	V32	6Q
885	V32	5A
886	V32	—
902	V35	Fig. 1
903	V35	6AL
904	V35	Fig. 3
905	V35	Fig. 6
906P1	V33	7AN
907	V35	Fig. 6
908	V35	7AN
908A	V35	7AN
909	V35	Fig. 6
910	V35	7AN
911	V35	7AN
912	V35	Fig. 8
913	V35	Fig. 1
914	V35	Fig. 12
930B	V43	3G
938	V46	4E
950	V20	5K
951	V19	4M
954	V25	5BB
955	V25	5BC
955	V39	5BC
956	V25	5BB
957	V25	5BD
958	V25	5BD
958A	V25	5BD
958A	V39	5BD
959	V25	5BE
967	V32	3G
975A	V37	4AT
991	V32	—
1003	V37	4R
1005	V37	5AQ
1006	V37	4C
1201	V25	8BN
1203	V25	4AH
1204	V25	8BO
1206	V17	8BV
1221	V18	6F
1223	V16	7R
1229	V20	4K
1230	V20	4D
1231	V17	8V
1232	V17	8V
1247	V29	—
1265	V32	4AJ
1266	V32	4AJ
1267	V32	4V
1273	V17	8V
1274	V37	6S
1275	V37	4C
1276	V25	4D
1280	V24	8V
1284	V24	8V
1291	V21	7BE
1293	V21	4AA
1294	V21	4AH
1299	V21	6BB
1602	V40	4D
1603	V18	6F
1608	V40	4D
1609	V25	5B
1610	V50	T-5C
1611	V14	7S
1612	V14	7T
1613	V50	7S
1614	51	7AC
1616	V37	4P
1619	V51	T-9H
1620	V14	7R
1621	V14	7S
1622	V14	7AC
1623	V41	3G
1624	V52	T-5D
1625	V52	5AZ
1626	V39	6Q
1627	V47	2N
1628	V41	T-4B
1629	V24	6RA
1631	V24	7AC
1632	V24	7AC
1633	V24	8BD
1634	V24	8S
1635	V16	8B
1641	V37	T-4A
1642	V17	7BH
1644	V24	Fig. 7
1654	V37	2Z
1800	V35	6AL
1801	V35	Fig. 1
1802P1	V33	11A
1803P4	V34	6AL
1804P4	V34	6AL
1805P1	V33	11A
1806P1	V33	11A
1809P4	V34	8BR
1811P1	V34	6AZ
1851	V14	7R
1852	V13	8N
1853	V13	8N
2001	V35	4AA
2002	V35	Fig. 1
2005	V35	Fig. 1
2050	V32	8BA
2051	V32	8BA
2523N/128AS	V32	5A
5514	V43	4BO
5516	V51	7CL
5517	V37	5BU
5556	V39	4D
5562	V53	Fig. 54
5590	V28	7BD
5591	V28	7BD
5618	V50	7CU
5633	V30	—
5634	V30	—
5637	V30	—
5638	V30	—
5639	V31	8DL
5640	V30	—
5641	V31	6CJ
5642	V31	—
5643	V31	8DD
5644	V31	4CN
5645	V31	—
5647	V31	B1
5648	V46	—
5651	V32	5BO
5654	V28	7BD
5663	V32	7CE
5670	V28	8CJ
5675	V39	Fig. 36
5679	V17	7CX
5686	V28	Fig. 29
5686	V50	Fig. 29
5687	V28	9H
5691	V16	8BD
5692	V16	8BD
5693	V14	8N
5718	V31	8DK
5719	V31	8DK
5722	V28	5CB
5725	V28	7CM
5726	V28	6BT
5749	V28	7BK
5750	V28	7CH
5751	V28	9A
5763	V50	9K
5764	V39	Fig. 36
5765	V39	Fig. 36
5766	V39	Fig. 36
5767	V39	Fig. 36
5768	V25	Fig. 36
5812	V50	7CQ
5812	V28	7CQ
5814	V29	9A
5823	V32	4CK
5824	V24	7S
5825	V38	4P
5840	V31	8DL
5866	V47	Fig. 5
5876	V39	Fig. 36
5879	V29	9AD

VACUUM-TUBE DATA

Type	Page	Base
5881	V51	7AC
5881	V16	7AC
5896	V31	8DJ
5897	V31	8DK
5898	V31	8DK
5899	V31	8DL
5900	V31	8DL
5901	V31	8DL
5902	V31	8DL
5903	V31	8DJ
5904	V31	8DK
5905	V31	8DL
5906	V31	8DL
5907	V31	8DL
5908	V31	8DC
5915	V29	7CH
5916	V31	8DC
5933	V52	5AW
5963	V29	9A
5964	V29	7BF
7000	V16	7R
7193	V39	4AM
7700	V18	6F
8000	V48	2N
8001	V53	7BM
8003	V46	3N
8005	V45	3G
8008	V38	Fig. 11
8010-R	V42	—
8012	V41	T-4BB
8013-A	V38	4P
8016	V38	4AC
8020	V38	4P
8025	V41	4AQ
9001	V29	7PM
9002	V29	7TM
9002	V39	7TM
9003	V29	7PM
9004	V25	4BJ
9005	V25	5BG
9006	V29	6BH
AT-340	V55	5BK
AX9900	V47	Fig. 5
AX9903	V52	Fig. 10
AX9905	V51	Fig. 34
BA	V36	4J
BH	V36	4J
BR	V36	4H
CE220	V36	4P
CK501	V29	—
CK502	V29	—
CK503	V29	—
CK504	V29	—
CK505	V29	—
CK506	V29	—
CK507	V29	—
CK509	V30	—
CK510	V30	—
CK512	V30	—
CK515BX	V30	—
CK520AX	V30	—
CK521AX	V30	—
CK522AX	V30	—
CK523AX	V30	—
CK524AX	V30	—
CK525AX	V30	—
CK526AX	V30	—
CK527AX	V30	—
CK529AX	V30	—
CK551AXA	V30	—
CK553AXA	V30	—
CK556AX	V30	—
CK568AX	V30	—
CK569AX	V30	—
CK605CX	V30	—
CK606BX	V30	—
CK608CX	V30	—
CK619CX	V30	—
CK624CX	V30	—
CK650AX	V30	—
CK703	V56	—
CK1005	V37	5AQ
CK1006	V37	4C
CK1007	V37	T-9G
CK1009	V37	—
CK5672	V30	—
DR3B27	V36	4P
DR123C	V46	Fig. 26
DR200	V47	2N
EF50	V25	9C
F123A	V46	Fig. 26
F127A	V48	Fig. 26
G84	V36	4B
GL2C44	V25	Fig. 17
GL2C44	V39	Fig. 17
GL5C24	V48	Fig. 26
GL5D24	V55	5BK
GL146	V47	T-4BG
GL152	V47	T-4BG
GL159	V49	T-4BG
GL169	V49	T-4BG
GL446A	V39	Fig. 19
GL446A	V25	Fig. 19
GL446B	V25	Fig. 19
GL446B	V39	Fig. 19
GL464A	V25	Fig. 17
GL464A	V39	Fig. 17
GL559	V25	Fig. 18
GL592	V48	Fig. 52
GL8012A	V41	T-4BB
HD203A	V47	3N
HF60	V44	2D
HF75	V44	2D
HF100	V44	2D
HF120	V45	4F
HF125	V45	—
HF130	V46	—
HF140	V45	4F
HF150	V46	—
HF175	V46	T-3AC
HF200	V47	2N
HF250	V47	2N
HF300	V48	2N
HK24	V41	3G
HK54	V42	2D
HK57	V53	5BK
HK154	V42	2D
HK158	V42	2D
HK252L	V47	4BC
HK253	V37	4AT
HK254	V45	2N
HK257	V54	7BM
HK257B	V54	7BM
HK304L	V49	4BC
HK354	V47	2N
HK354C	V47	2N
HK354D	V47	2N
HK354E	V47	2N
HK354F	V47	2N
HK454H	V49	2N
HK454L	V49	2N
HK654	V49	2N
HV12	V48	3N
HV18	V47	2N
HV27	V48	3N
HY6J5GTX	V39	6Q
HY6L6GTX	V51	7AC
HY6V6GTX	V51	7AC
HY24	V39	4D
HY25	V41	3G
HY30Z	V41	4BO
HY31Z	V41	T-4D
HY40	V42	3G
HY40Z	V42	3G
HY51A	V43	3G
HY51B	V43	3G
HY51Z	V43	4BO
HY57	V42	3G
HY60	V51	5AW
HY61	V52	5AW
HY63	V50	T-8DB
HY65	V51	T-8DB
HY67	V53	T-5DB
HY69	V52	T-5D
HY75	V40	2T
HY75A	V40	2T
HY113	V30	5K
HY114B	V39	2T
HY115	V30	5K
HY123	V30	5K
HY125	V30	5K
HY145	V30	5K
HY155	V30	5K
HY615	V39	T-8AG
HY801A	V40	4D
HY866jr	V37	4P
HY1231Z	V41	T-4D
HY1269	V53	T-5DB
HYE1148	V39	T-8AG
KY21	V32	—
KY866	V32	Fig. 8
M54	V30	—
M64	V30	—
M74	V30	—
NU-2C35	V25	Fig. 38
PE340	V54	5BK
QK159	V56	Fig. 63
RK2J22	V57	—
RK2J23	V57	—
RK2J24	V57	—
RK2J25	V57	—
RK2J26	V57	—
RK2J27	V57	—
RK2J28	V57	—
RK2J29	V57	—
RK2J30	V57	—
RK2J31	V57	—
RK2J32	V57	—
RK2J33	V57	—
RK2J34	V57	—
RK2J36	V57	—
RK2J38	V57	—
RK2J39	V57	—
RK2J48	V57	—
RK2J49	V57	—
RK2J50	V57	—
RK2J54	V57	—
RK2J55	V57	—
RK2J56	V57	—
RK2J58	V57	—
RK2J61A	V57	—
RK2J62A	V57	—
RK2J66	V57	—
RK2J67	V57	—
RK2J68	V57	—
RK2J69	V57	—
RK4J31	V57	—
RK4J32	V57	—
RK4J33	V57	—
RK4J34	V57	—
RK4J35	V57	—
RK4J36	V57	—
RK4J37	V57	—
RK4J38	V57	—
RK4J39	V57	—
RK4J40	V57	—
RK4J41	V57	—
RK4J43	V57	—
RK4J44	V57	—
RK4J53	V57	—
RK4J54	V57	—
RK4J55	V57	—
RK4J56	V57	—
RK4J57	V57	—
RK4J58	V57	—
RK4J59	V57	—
RK725A	V57	—
RK10	V40	4D
RK11	V41	3G
RK12	V41	3G
RK15	V19	4D
RK16	V19	5A
RK17	V19	5F
RK18	V41	3G
RK19	V38	4AT
RK20	V52	T-5C
RK20A	V52	T-5C
RK21	V38	4P
RK22	V38	T-4AG
RK23	V50	6BM
RK24	V39	4D
RK24	V20	4D
RK25	V50	6BM
RK25B	V50	6BM
RK28	V54	5J
RK28A	V54	5J
RK30	V41	2D
RK31	V41	3G
RK32	V42	2D
RK33	V39	T-7DA
RK34	V39	T-7DC
RK35	V42	2D
RK36	V45	2D
RK37	V42	2D
RK38	V45	2D
RK39	V51	5AW
RK41	V51	5AW
RK42	V21	4D
RK43	V21	6C
RK44	V50	6BM
RK46	V52	T-5C
RK47	V53	T-5D
RK48	V54	T-5D
RK48A	V54	T-5D
RK49	V51	6A
RK51	V43	3G
RK52	V43	3G
RK56	V50	5AW
RK57	V46	3N
RK58	V45	3N
RK59	V40	T-4D
RK60	V37	T-4AG
RK61	V30	—
RK62	V32	4D
RK63	V48	2N
RK63A	V48	2N
RK64	V50	5AW
RK65	V55	T-3BC
RK66	V52	T-5C
RK75	V51	T-5C
RK100	V40	6A
RK705A	V37	T-3AA
RK866	V37	4P
RM208	V32	—
RM209	V32	—
SD917A	V30	—
SD828A	V30	—
SD828E	V30	—
SD1103	V56	—
SD1104	V56	—
SN944	V30	—
SN946	V30	—
SN947D	V30	—
SN948C	V30	—
SN953D	V30	—
SN954	V30	—
SN955B	V30	—
SN956B	V31	—
SN957A	V31	—
SN1006	V31	—
SN1007B	V31	—
T20	V40	3G
T21	V51	6A
T40	V42	3G
T55	V43	3G
T60	V43	2D
T100	V44	2D
T125	V46	2N
T200	V48	2N
T300	V48	—
T814	V48	3N
T822	V48	3N
TB35	V52	Fig. 54
TUF20	V40	2T
TW75	V44	2D
TW150	V47	2N
TZ20	V40	3G
TZ40	V42	3G
UE100	V44	2D
UE468	V47	Fig. 57
UH35	V44	3G
UH50	V42	2D
UH51	V42	2D
V70	V44	3N
V70A	V44	3N
V70B	V44	3G
V70C	V44	3G
V70D	V45	3G
VR75	V32	4AJ
VR90	V32	4AJ
VR105	V32	4AJ
VR150	V32	4AJ
VT52	V25	4D
VT127A	V45	T-4B
VT191	V41	—
WE304A	V42	2D
X6030	V25	Fig. 4
XXB	V25	Fig. 9
XXD	V24	8AC
XXL	V17	5AC
XXFM	V25	8BZ
Z225	V37	4P
Z668	V56	—
ZB60	V44	2D
ZB120	V44	4E

CHAPTER 27

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:

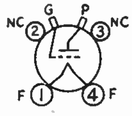
A = Anode
B = Beam
BP = Bayonet Pin
BS = Base sleeve
D = Deflecting Plate
F = Filament
FE = Focus Elect.
G = Grid
H = Heater
IC = Internal Connection
IS = Internal Shield
K = Cathode
NC = No Connection
P = Plate (Anode)
P_1 = Starter-Anode
P_{BF} = Beam-Forming Plates
RC = Ray-Control Electrode
Ref = Reflector
S = Shell
TA = Target
● = Gas-Type Tube
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.

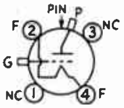
Generally when the No. 1 pin of a metal-type tube in Table I, with the exception of all triodes, is shown connected to the shell, the No. 1 pin in the glass (G or GT) equivalent is connected to an internal shield.

R.T.M.A. TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are shown above.



2D



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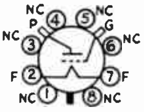

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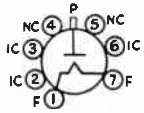

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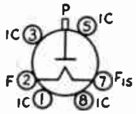

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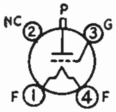

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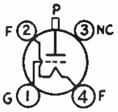

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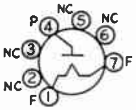

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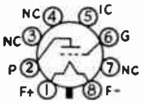

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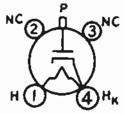

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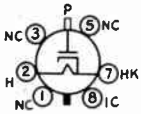

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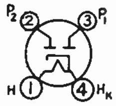

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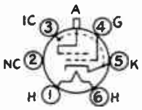

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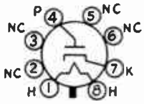

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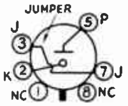

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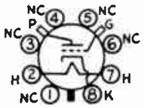

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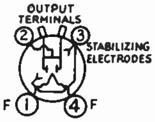

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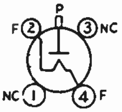

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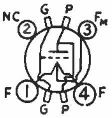

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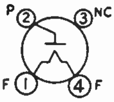
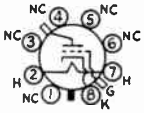


4B



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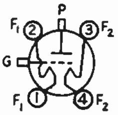

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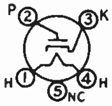

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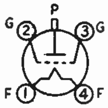

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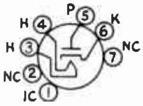

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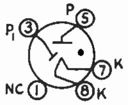

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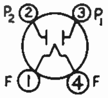

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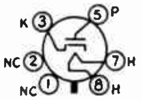

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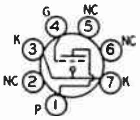

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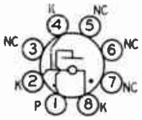

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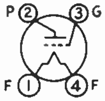

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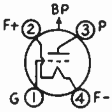

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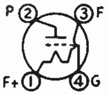

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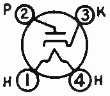

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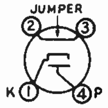

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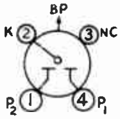

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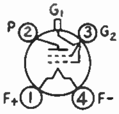

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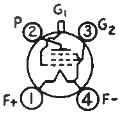

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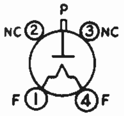

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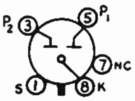

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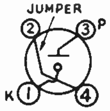

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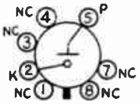

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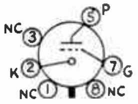

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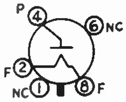

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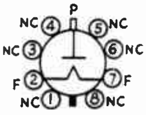

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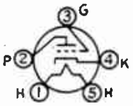

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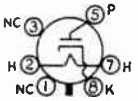

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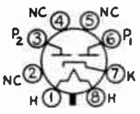
R.T.M.A. TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page V5.



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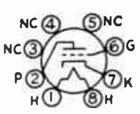

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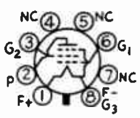

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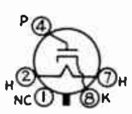

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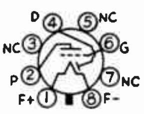

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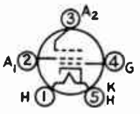

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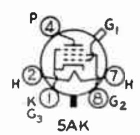

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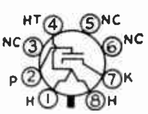

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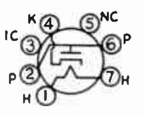

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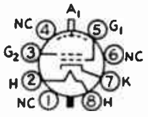

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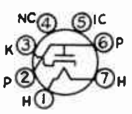

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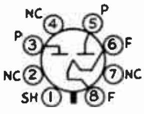

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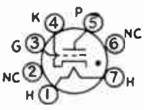

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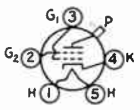

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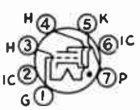

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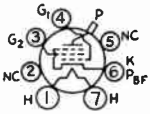

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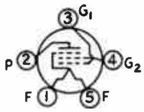

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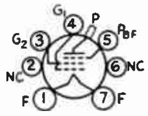

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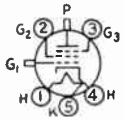

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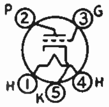

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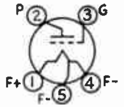

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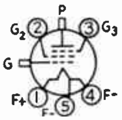

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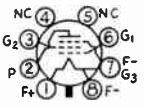

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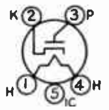

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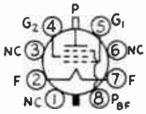

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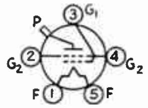

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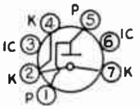

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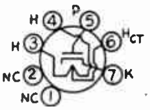

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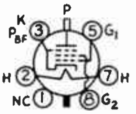

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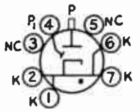

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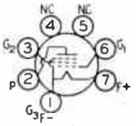

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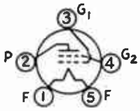

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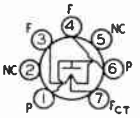

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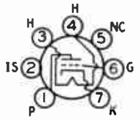

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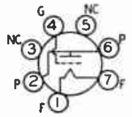

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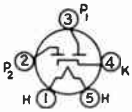

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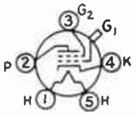

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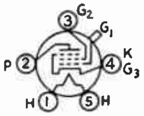

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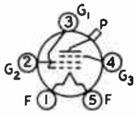

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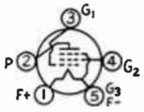

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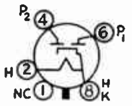

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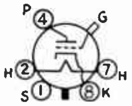

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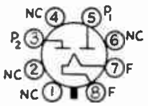

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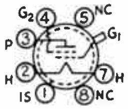

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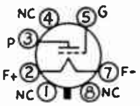

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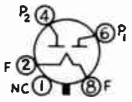

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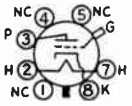

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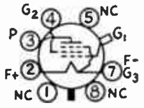

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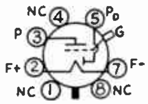

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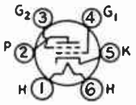

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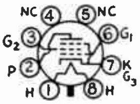

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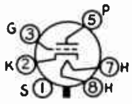

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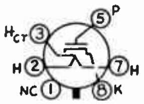

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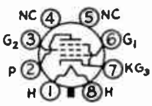

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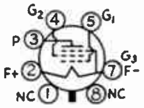

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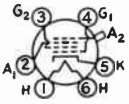

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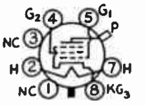

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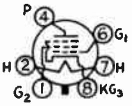

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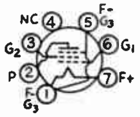

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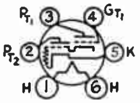

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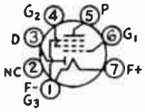
R.T.M.A. TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page V5.



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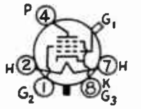

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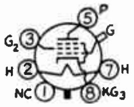

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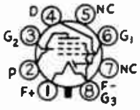

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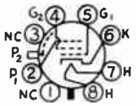

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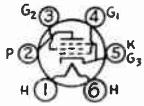

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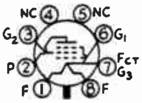

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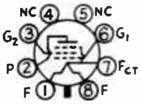

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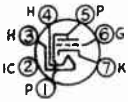

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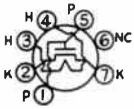

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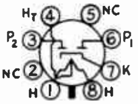

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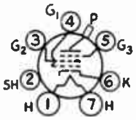

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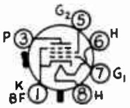

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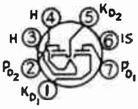

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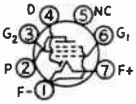

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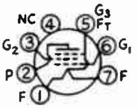

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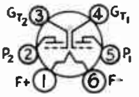

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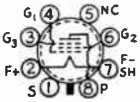

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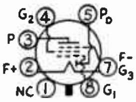

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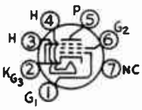

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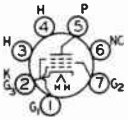

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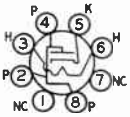

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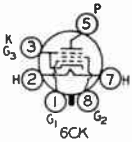

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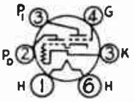

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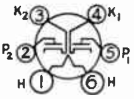

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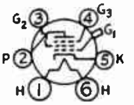

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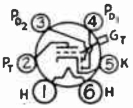

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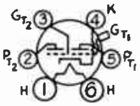

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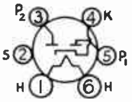

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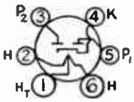

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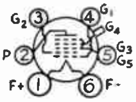

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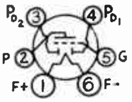

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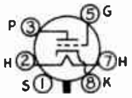

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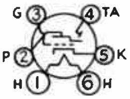

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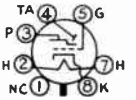

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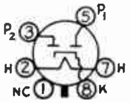

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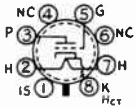

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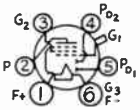

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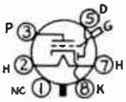

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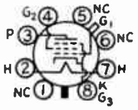

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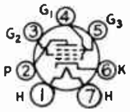

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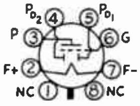

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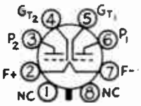

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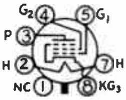

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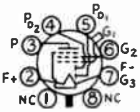

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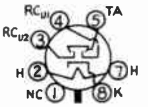

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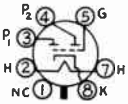

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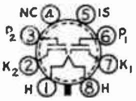

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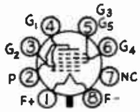

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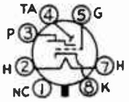

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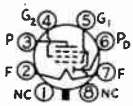

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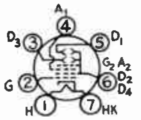

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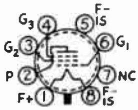

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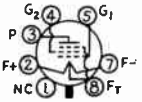

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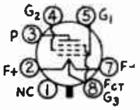

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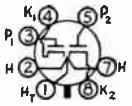

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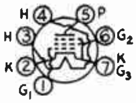

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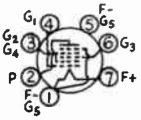

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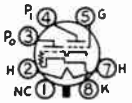

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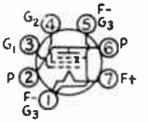
R.T.M.A. TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page V5.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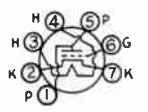
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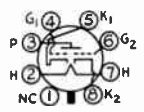
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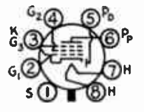
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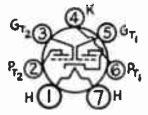
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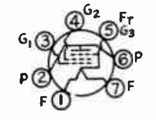
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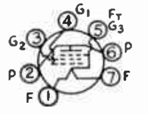
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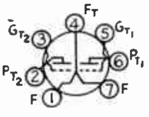
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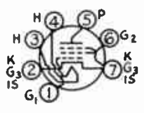
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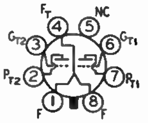
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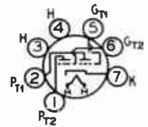
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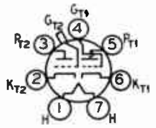
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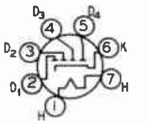
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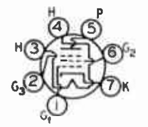
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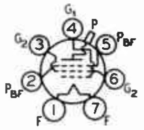
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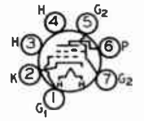
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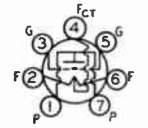
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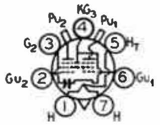
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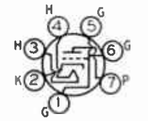
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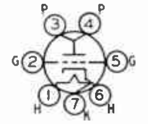
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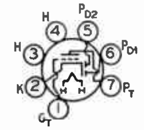
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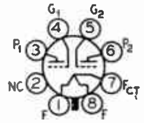
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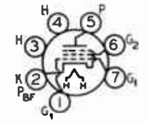
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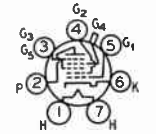
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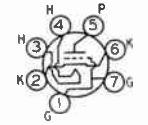
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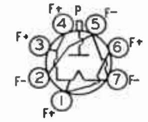
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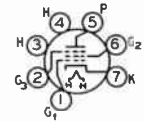
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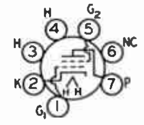
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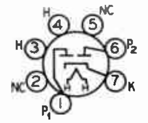
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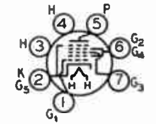
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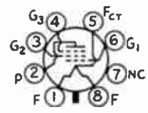
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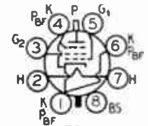
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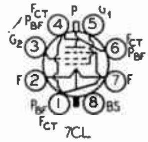
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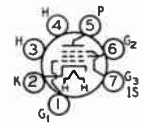
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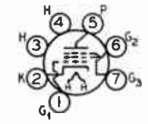
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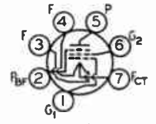
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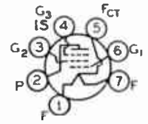
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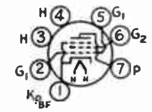
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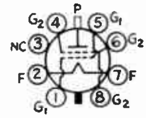
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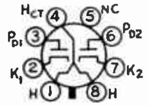
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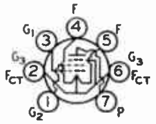
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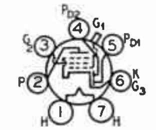
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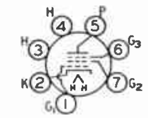
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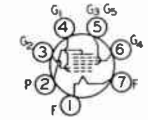
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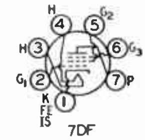
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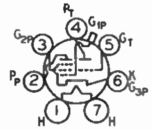
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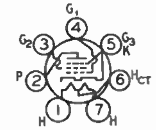
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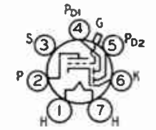
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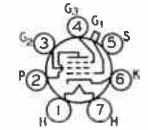
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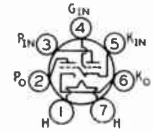
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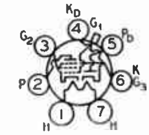
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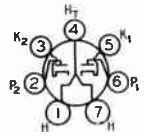
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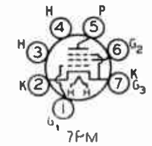
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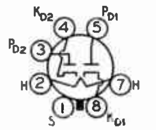
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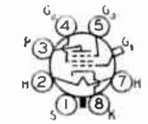
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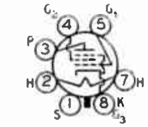
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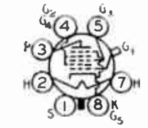
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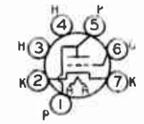
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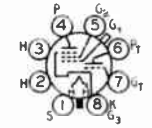
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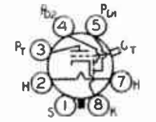
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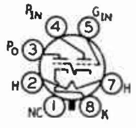
R.T.M.A. TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page V5.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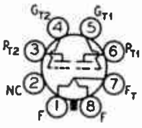
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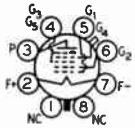
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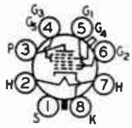
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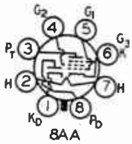
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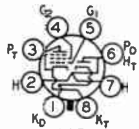
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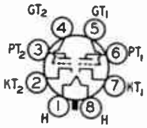
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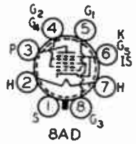
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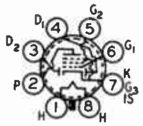
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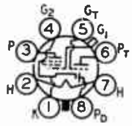
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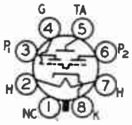
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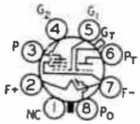
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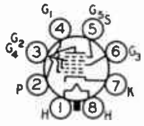
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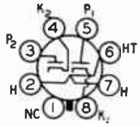
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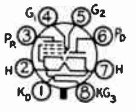
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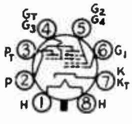
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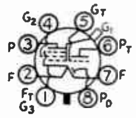
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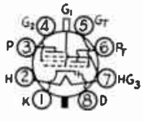
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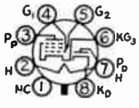
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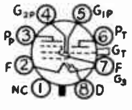
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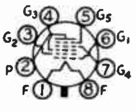
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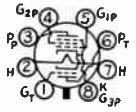
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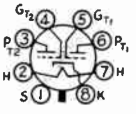
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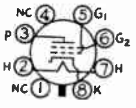
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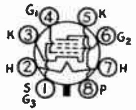
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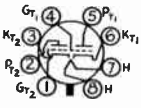
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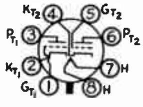
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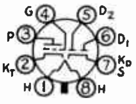
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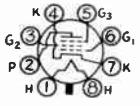
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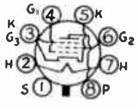
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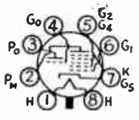
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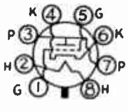
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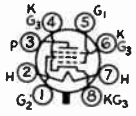
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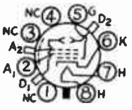
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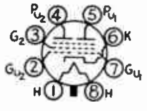
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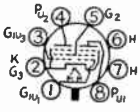
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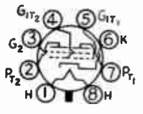
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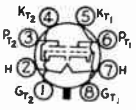
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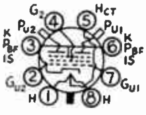
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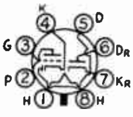
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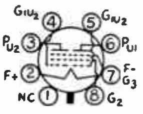
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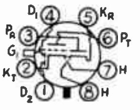
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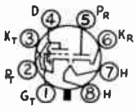
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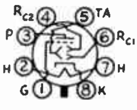
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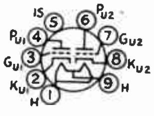
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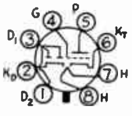
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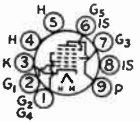
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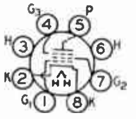
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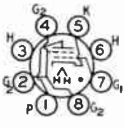
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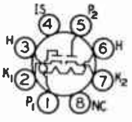
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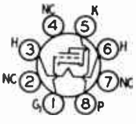
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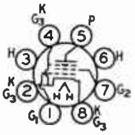
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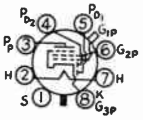
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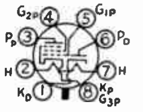
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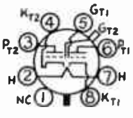
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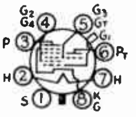
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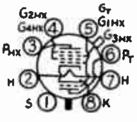
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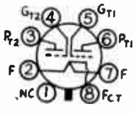
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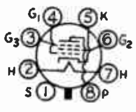
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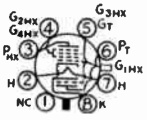
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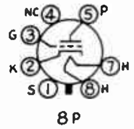
R.T.M.A. TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page V5.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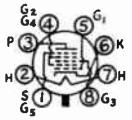
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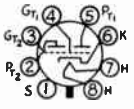
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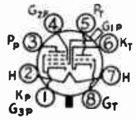
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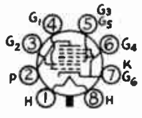
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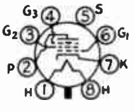
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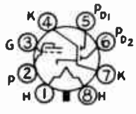
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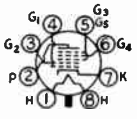
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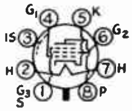
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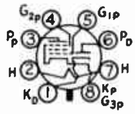
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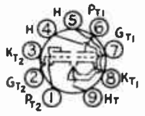
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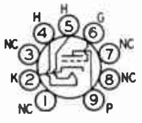
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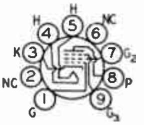
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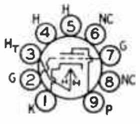
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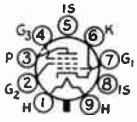
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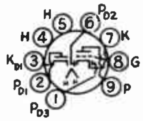
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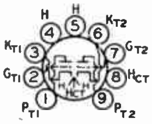
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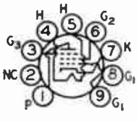
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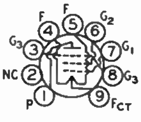
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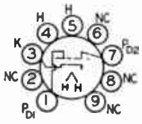
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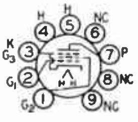
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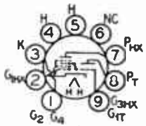
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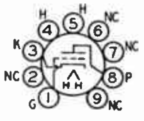
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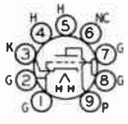
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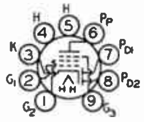
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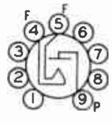
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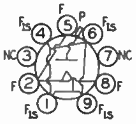
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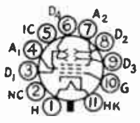
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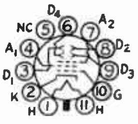
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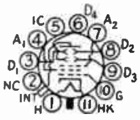
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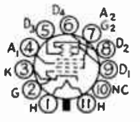
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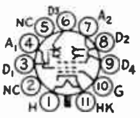
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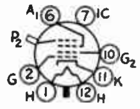
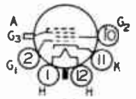
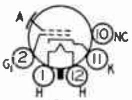

12 C


12D


12 E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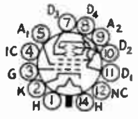
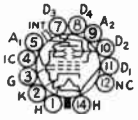
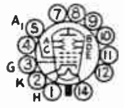
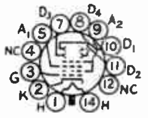

14 A


14 B


14 F


14 G


B1

SUPPLEMENTARY BASE DIAGRAMS

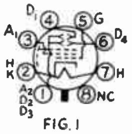


FIG. I

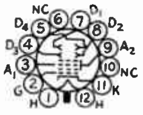


FIG. 2

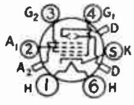


FIG. 3

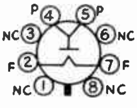


FIG. 4

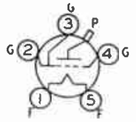


FIG.5

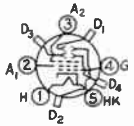


FIG 6

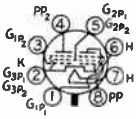


FIG. 7

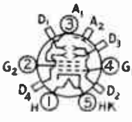


FIG. 8

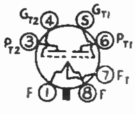


FIG. 9

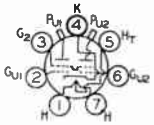


FIG. 10

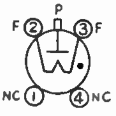


FIG. II

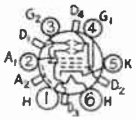


FIG. 12

SUPPLEMENTARY BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page V5.

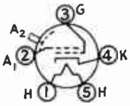


FIG. 13

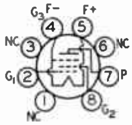


FIG. 14

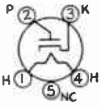


FIG 15

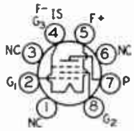


FIG. 16

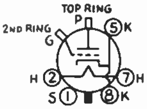


FIG. 17

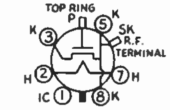


FIG. 18

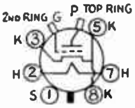


FIG. 19

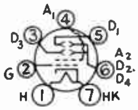


FIG. 20

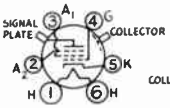


FIG. 21

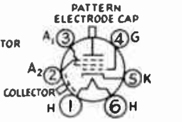


FIG. 22

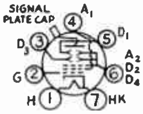


FIG 23

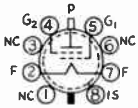


FIG. 24

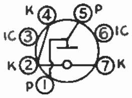


FIG. 25

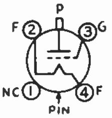


FIG. 26

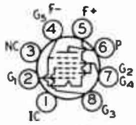


FIG. 27

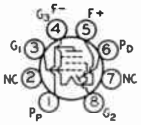


FIG. 28

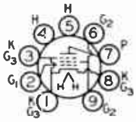


FIG. 29

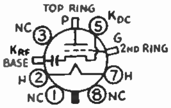


FIG. 30

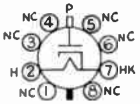


FIG 31

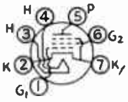


FIG. 32

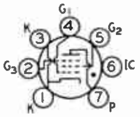


FIG. 33

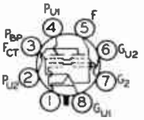


FIG. 34

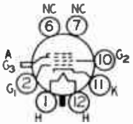


FIG. 35

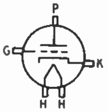


FIG. 36

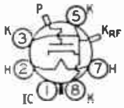


FIG. 37

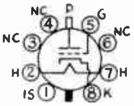


FIG. 38

FIG. 39

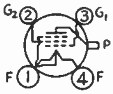


FIG. 40

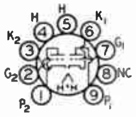


FIG. 41

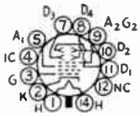


FIG. 47

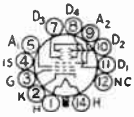


FIG 49

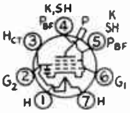


FIG. 50

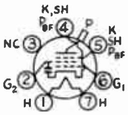


FIG. 51

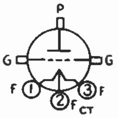


FIG. 52

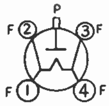


FIG. 53

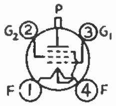


FIG. 54

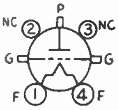


FIG. 56

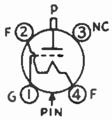


FIG. 57

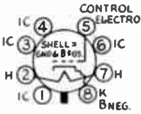


FIG. 58

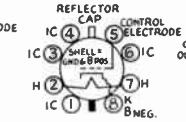


FIG 59

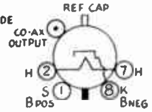


FIG. 60

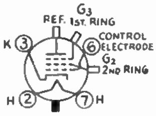


FIG. 61

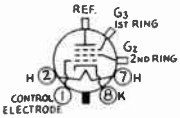


FIG 62

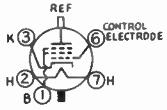
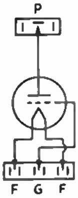


FIG. 63

SUPPLEMENTARY "T"-GROUP BASE DIAGRAMS

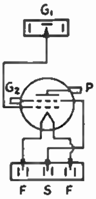
Bottom views are shown. Terminal designations on sockets are given on page V5.



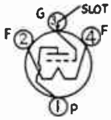
T-1A

T-1AA

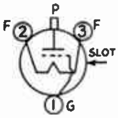
T-1AB



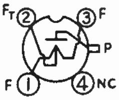
T-1B



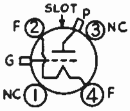
T-2A



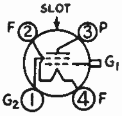
T-2AA



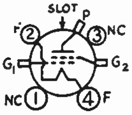
T-3AA



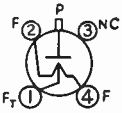
T-3AC



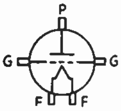
T-3B



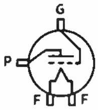
T-3BC



T-4A



T-4AD



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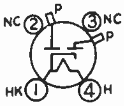

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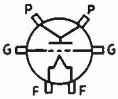

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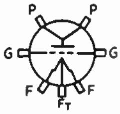

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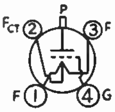

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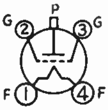

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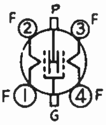

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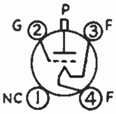

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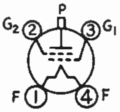

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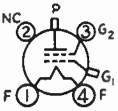

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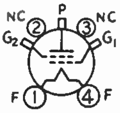

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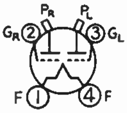

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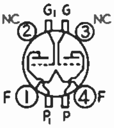

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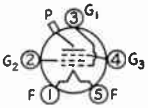

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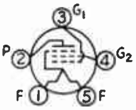

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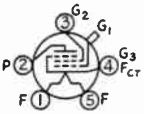

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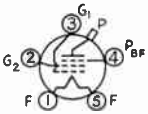

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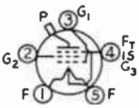

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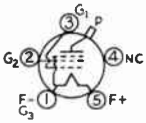

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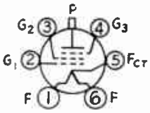

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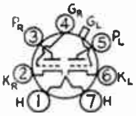

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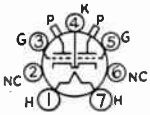

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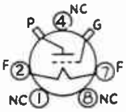

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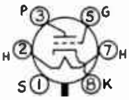

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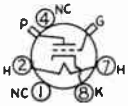

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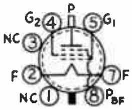

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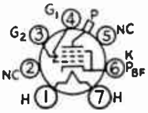

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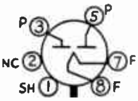

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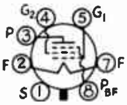

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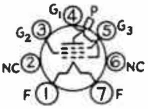

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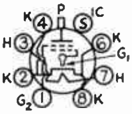


T-9J

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 $\mu\mu$fd.			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 Converter	8A	6.3	0.3	Osc. Grid leak = 50000Ω			Converter	250	−3.0	100	2.7	3.5	Anode-grid (No. 2) 250 volts max. thru 20,000 ohms					6A8
6AB7 1853	Remote Cutoff 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	Sharp Cutoff 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 of 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$_1$^{6}	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
6SB7Y	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	—	—	—	—	—	
6SC7	Twin-Triode	85	6.3	0.3	—	—	—	Class-A Amp.	250	−2.0	—	—	2.0	53000	1325	70	—	—	6SC7

TABLE I—METAL RECEIVING TUBES—*Continued*

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												
6SF5	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
6SF7	Diode 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
6SG7	Semivariable-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
6SH7	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
6SJ7[4]	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
6SK7	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	BQ	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.[6] [1]	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—Scrn 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												
2B22	Diode	Fig. 37	6.3	0.75	2.2	—	—	U.h.f. Detector		Average cathode Ma. = 5; Output volts = 50 d.c.; Load resistance = 10000Ω.									2B22
2C22	Triode	4AM	6.3	0.3	2.2	0.7	3.60	Class-A Amp.	300	−10.5	—	—	11	6600	3000	20	—	—	2C22
6A5GT	Triode Power Amplifier	6T	6.3	1.0	—	—	—	Class-A Amp.[4]	250	−45.0	—	—	60	800		4.2	2500	3.75	6A5GT
								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						
6AC5GT	High-μ Power-Amplifier Triode	6Q	6.3	0.4	—	—	—	P.P. Class B[5]	250	0	—	—	5.0	36700	3400	125	10000[6]	8.0	6AC5GT
								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 far 135°; 45 for 0°. Target current 1.5 ma. for 0°.								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

TABLE II—6.3-VOLT GLASS TUBES WITH OCTAL BASES—*Continued*

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												
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			
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[10]	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
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	
6AU5GT	Beam Pentode	6CK	6.3	1.25	11.3	7	0.5	Horz. Def. Amp.	450[11]	−50[11]	—	—	100[11]	Peak pos. plate pulse = 5000 volts.					6AU5GT
6AV5GT	Beam Pentode	6CK	6.3	1.2	—	—	—	Horz. Def. Amp.	500[11]	−50[11]	175[11]	—	100[11]	Peak pos. plate pulse = 4500 volts.					6AV5GT
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
6BD5GT	Beam Pentode	6CK	6.3	0.9	—	—	—	Horz. Def. Amp.	325[11]	—	325[11]	—	100[11]	Peak pos. plate pulse = 4000 volts.					6BD5GT
6BQ6GT	Beam Pentode	6AM	6.3	1.2	—	—	—	Deflection Amp.	550[11]	—	150	—	100[11]	Peak pos. plate pulse = 4000 volts.					6BQ6GT
6BG6	Beam Power Amplifier	5BT	6.3	0.9	11	6.5	0.5	Deflection Amp.	700[11]	−50[11]	350	—	100[11]	Peak pos. plate pulse = 6000 volts.					6BG6
6C8G	Twin Triode	8G	6.3	0.3	—	—	—	Amp. 1 Section	250	−4.5	—	—	3.1	26000	1450	38	—	—	6C8G
6CD6G	Beam Pentode	5BT	6.3	2.5	26	10	1.0	Horz. Def. Amp.	700[11]	−50[11]	175[11]	—	170[11]	Peak pos. plate pulse = 6000 volts.					6CD6G
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[10]	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

TABLE II—6.3-VOLT GLASS TUBES WITH OCTAL BASES—*Continued*

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												
6SH7L	Pentode R.F. Amp.	8BK	6.3	0.3	—	—	—	Class-A Amplifier	250	− 1.0	150	4.1	10.8	900000	4900	—	—	—	6SH7L
6SL7GT	Twin Triode	8BD	6.3	0.3	—	—	—	Class-A Amplifier	250	− 2.0	—	—	2.3[1]	44000	1600	70	—	—	6SL7GT
6SN7GT 6SN7GTA	Twin Triode	8BD	6.3	0.6	—	—	—	Class-A Amplifier	250	− 8.0	—	—	9.0[1]	7700	2600	20	—	—	6SN7GT 6SN7GTA
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	Pentade 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
5881	Beam Power Amp.	7AC	6.3	0.9	—	—	—	Audio Amplifier	Characteristics same as 6L6, Table I										5881
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.
[11] Max. value.

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		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												
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-A₁ 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	6.6	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-A₁ 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-A₁ 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-A₁ Amplifier	250	250*	250	1.9	6.8	1000000	3300	—	—	—	7AH7
7AJ7	Sharp Cut-off Pentode	8V	6.3	0.3	6.0	6.5	0.007	Class-A₁ Amp.	250	− 3	100	0.7	2.2	1 Meg.	1575	—	—	—	7AJ7
									100	− 1	100	1.8	5.5	400000	2275	—	—	—	
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-A₁ 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

TABLE III—7-VOLT LOCK-IN-BASE TUBES—*Continued*

Type	Name	Socket Connections	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												
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
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	− 2.5	—	—	10.0	10400	5000	—	—	—	7F8
									180	− 1.0			12.0	8500	7000				
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	− 3.0	100	0.7	2.2	1000000	1575	—	—	—	1273
									100	− 1.0	100	1.8	5.7	400000	2275	—	—	—	
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 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												
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

TABLE IV—6.3-VOLT GLASS RECEIVING TUBES—*Continued*

Type	Name	Base	Socket Connections	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												
6B5	Direct-Coupled Power Amplifier	M.	6AS	6.3	0.8	—	—	—	Class-A Amp.[9]	300	0	—	6[1]	45	241000	2400	58	7000	4.0	6B5
									Push-Pull Amp.[10]	400	−13.0	—	4.5[1]	40	—	—	—	10000[11]	20	
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	Cut-off Grid Bias = −22 v.			0.24	Target Current 4 ma.				—	6U5/6G5
										100	Cut-off Grid Bias = −8 v.			0.19	Target Current 1 ma.					
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	1.7	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 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												
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 2A3											RK17

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 Connections	Filament		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 Outpu Watts	Type
				Volts	Amp.	In	Out	Plate-Grid												
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		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												
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		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												
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
1J6GT	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		Capacitance $\mu\mu$fd.			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
				Volts	Amp.	In	Out	Plate-Grid												
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	Osc. Grid leak 200000Ω			Converter	90	0	45	0.7	0.6	600000	250	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	0	90	0.8	3.5	275000	1100	—	—	—	1AB5
										150	−1.5	150	2.0	6.8	125000	1350				
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.	90	0	—	—	1.0	45000	675	30	—	—	1G6GT
									Class-B Amp.	90	0	—	—	1/7	34 volts input per grid			12000	675	
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	Osc. Grid leak 200000Ω			Converter	90	0	45	0.6	0.55	750000	250	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	Osc. Grid leak 200000Ω			Converter	90	0	35[1]	0.7	0.75	650000	275	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 #	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.	90	0	67.5	0.38	1.45	700000	665	—	—	—	1SB6GT
									R.C. Amplifier	90	0	90	Screen resistor 5 meg., grid 10 meg.					1 meg.	110[2]	
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 Connec-tions	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												
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.	100	− 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										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

TABLE IX—HIGH-VOLTAGE HEATER TUBES—*Continued*

Type	Name	Base	Socket Connections	Heater		Capacitance $\mu\mu$fd.			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												
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	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	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	
25AV5GT	Beam Pentode	O.	6CK	25	0.3	—	—	—	Horz. Def. Amp.	250[9]	−50[9]	175[9]	—	100[9]	Peak pos. plate pulse =4500 volts.					25AV5GT
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
35L6GT	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	35L6GT
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		Capacitance μμfd.			Uso	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												
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 II										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 [2] 6000 [4]	0.08 [2] 0.175 [3]	28D7
5824	Pentode	O.	7S	25	0.3	—	—	—	Class-A_1 Amp.	135	–22	135	2.5/14.5	61/69	15000	5000	—	1700	4.3	5824

* 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.
[9] Max. value.

TABLE X—SPECIAL RECEIVING TUBES

Type	Name	Base	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												
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	Beam Pentode	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		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												
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] [7]	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
5768	U.h.f. "Rocket" Triode	N.	Fig. 36	6.3	0.4	1.2	0.01	1.3	1000–3000-Mc. Amplifier	250	− 1	—	—	9.3	—	4500	85	—	—	5768
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
XXB	Twin-Triode Frequency Converter	L.	Fig. 9	2.8/ 1.4	0.05/ 0.10	—	—	—	Converter[1]	90[1]	0	—	—	4.5[4] 4.5[5]	11200[4] 11200[5]	1300[4] 1300[5]	14.5[5]	—	—	XXB
				3.2[6]/ 1.6	— —						− 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.

[6] Dry battery operation.

[5] Section No. 2.

[7] Discontinued.

TABLE XI—MINIATURE RECEIVING TUBES

Other miniature types in Tables XIII and XV

Type	Name	Base	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[4]	Load Resistance Ohms	Power Output Watts	Prototype
				Volts	Amp.	In	Out	Plate-Grid												
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.					—
1AF4	Pentode	B.	6AR	1.4	0.025	3.8	7.6	.008	Class-A$_1$ Amp.	90	0	90	0.5	1.65	1800000	950	—	—	—	—
1AF5	Diode Pentode	B.	6AU	1.4	0.025	—	—	—	Class-A$_1$ Amp.	90	0	90	0.4	1.1	2000000	600	—	—	—	—
1C3	Triode	B.	5CF	1.4	0.05	0.9	4.2	1.8	Class-A$_1$ Amp.	90	− 3	—	—	1.4	19000	760	14.5	—	—	1LE3
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
1L6	Pentagrid Converter	B.	7DC	1.4	0.05	7.5	12	0.3	Converter	90	0	45	0.6	0.5	650000	300	—	—	—	1LA6
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	—	—	—	—
1U6	Pentagrid Converter	B.	7DC	1.4	0.025	8	12	0.4	Converter	90	0	45	0.55	0.55	600000	275	—	—	—	—
1W4	Power Amplifier Pentode	B.	5BZ	1.4	0.05	3.6	7	0.1	Class-A$_1$ Amp.	90	− 9	90	1	5	300000	925	—	12000	0.2	1LB4
2C51	Twin Triode	B.	8CJ	6.3	0.3	2.2	1.0	1.3	Class-A$_1$ 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-A$_1$ Single	250	450*	250	7.4[2]	44[2]	63000	3700	40[5]	4500	4.5	—
									Class-A$_1$ Amp.[3]	250	225*	250	14.8[2]	88[2]	—	—	80[5]	9000[6]	9	
									Class-AB$_1$ Amp[3]	250	−25	250	13.5[2]	80[2]	—	—	48[5]	8000[6]	12.5	
									Class-AB$_2$ Amp.[3]	250	−30	250	20[2]	120[2]	—	—	40[5]	3800[6]	17	
3A4	Power Amplifier Pentode	B.	7BB	1.4	0.2	4.8	4.2	0.34	Class-A$_1$ Amp.	135	− 7.5	90	2.6	14.9[2]	90000	1900	—	8000	0.6	—
				2.8	0.1					150	− 8.4	90	2.2	14.1[2]	100000				0.7	
3A5	H.F. Twin Triode	B.	7BC	1.4	0.22	0.9	1.0	3.20	Class-A Amp.	90	− 2.5	—	—	3.7	8300	1800	15	—	—	—
				2.8	0.11															
3E5	Power Amplifier Pentode	B.	6BX	1.4	0.05	—	—	—	Class-A$_1$ Amp.	90	− 8	90	1.5	5.5	120000	1100	—	8000	.175	—
				2.8	.025															
3Q4	Power Amplifier Pentode	B.	7BA	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							1.7	7.7	120000	2000			0.24	
3S4	Power Amplifier Pentode	B.	7BA	1.4	0.1	Parallel Filaments			Class-A Amp.	90	− 7.0	67.5	1.4	7.4	100000	1575	—	8000	0.27	3Q5GT
				2.8	0.05	Series Filaments							1.1	6.1		1425			0.235	
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	U.h.f. Triode	B.	5CE	6.3	0.15	2.2	0.5	1.5	Class-A Amp.	250	200*	—	—	10	10900	5500	60	—	—	Single unit 12AT7
6AG5	Sharp Cut-off Pentode	B.	7BD	6.3	0.3	—	—	—	Class-A Amp.	250	200*	150	2.0	7.0	800000	5000	—	—	—	6SH7GT
										100	100*	100	1.6	5.5	300000	4750	—	—		
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.[7]	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.[8]	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	—	—	
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
6A6M5	Power Amplifier Pentode	B.	6CH	6.3	0.2	—	—	—	Class-A$_1$ Amp.	250	−13.5	250	2.4	16	130000	2600	—	16000	1.4	—
6AM6	Pentode	B.	7DB	6.3	0.3	7.5	3.25	0.01	Class-A$_1$ Amp.	250	− 2	250	2.5	10	1000000	7500	—	—	—	—
6AN5	Power Amp. Pentode	B.	7BD	6.3	0.5	9.0	4.8	0.05	Class-A$_1$ Amp.	120	− 6	120	12	35	12500	8000	—	—	—	6AG7

TABLE XI—MINIATURE RECEIVING TUBES—*Continued*

Type	Name	Base	Socket Connections[1]	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[4]	Load Resistance Ohms	Power Output Watts	Prototype
				Volts	Amp.	In	Out	Plate-Grid												
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.										—
6AN7	Triode Hexode	B.	9Q	6.3	0.23	3.8	9.2	0.1	Converter	250	−2	85	3	3	—	750	—	—	—	—
6AQ5	Beam Power Tetrode	B.	7BZ	6.3	0.45	7.6	6.0	0.35	Class-A$_1$ Amp.	180	−8.5	180	4.0[2]	30[2]	58000	3700	29[4]	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-A$_1$ 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-A$_1$ 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	5.2	—	3200	—	—	—	—
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-A$_1$ 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	950	—	—	—	6SB7Y
6BC5	Pentode	B.	7BD	6.3	0.3	6.6	3.1	.02	Class-A$_1$ Amp.	250	180*	150	1.4	4.7	600000	4900	—	—	—	—
6BD6	Remote Cut-off Pentade	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
6BF5	Beam Power Pentode	B.	7BZ	6.3	1.2	—	—	—	Class-A$_1$ Amp.	110	−7.5	110	4.0/8.5	49/50	10000	7500	—	2500	1.9	—
6BF6	Duplex-Diode Triode	B.	7BT	6.3	0.3	1.8	1.1	2.0	Class-A$_1$ 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-A$_1$ 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-A$_1$ Amp.	250	−1	100	3.3	9.2	1300000	3800	—	—	—	6SS7GT
6BK6	Duodiode Triode	B.	7BT	6.3	0.3	—	—	—	Class-A$_1$ Amp.	250	−2	—	—	1.2	80000	1250	100	—	—	—
6BN6	Gated-beam Disc.	B.	7DF	6.3	0.3	4.2	3.3	.004	FM Disc.	80	−1.3	60	5	0.23	—	—	—	68000	—	—
6BN7	Dual Triode	B.	Fig. 41	6.3	0.75	5.5[7]	1.6[7]	3[7]	Class-A$_1$ Amp.[7]	250	−15	—	—	24	2200	5500	12	—	—	—
						1.4[8]	0.3[8]	0.7[8]	Class-A$_1$ Amp.[8]	120	−1	—	—	5	14000	2000	28	—	—	
6BT6	Duodiode Triode	B.	7BT	6.3	0.03	—	—	—	Class-A$_1$ Amp.	250	−3	—	—	1	58000	1200	70	—	—	—
6BU6	Duodiode Triode	B.	7BT	6.3	0.3	—	—	—	Class-A$_1$ Amp.	250	−9	—	—	9.5	8500	1900	16	10000	0.3	—
6C4	Triode Amplifier	B.	6BG	6.3	0.15	1.8	1.3	1.60	Class-A$_1$ Amp.	250	−8.5	—	—	10.5	7700	2200	17	—	—	6J5GT
6CB6	Sharp Cut-off Pentode	B.	7CM	6.3	0.3	6.3	1.9	0.02	Class-A$_1$ Amp.	200	180*	150	2.8	9.5	600000	6200	—	—	—	—
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-A$_1$ Amp.	100	100*	—	—	10.0	5000	11000	55	—	—	
6J6	Twin Triode	B.	7BF	6.3	0.45	2.2	0.4	1.6	Class-A$_1$ Amp. Mixer, Oscillator	100	50*	—	—	8.5	7100	5300	38	—	—	—
6M5	Power Amplifier Pentode	B.	9N	6.3	0.71	10	6.2	1	Class-A$_2$ Amp.	250	170*	250	5.2	36	40000	10000	—	7000	3.9	—
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	—	—	—
6N8	Duodiode Pentode	B.	9T	6.3	0.3	4	4.6	.002	Class-A$_1$ Amp.	250	−2	85	—	1.75	1600000	2200	—	—	—	—
6Q4	Grnd.-Grid Triode	B.	9S	6.3	0.48	5.4	.06	3.4	Class-A$_1$ Amp.	250	−1.5	—	—	15	—	12000	80	—	—	—
6R4	U.h.f. Triode	B.	9R	6.3	0.2	1.7	0.5	1.5	Class-A$_1$ Amp.	150	−2	—	—	30	—	5500	16	—	—	—
6R8	Triple Diode Triode	B.	9E	6.3	0.45	1.5	1.1	2.4	Class-A$_1$ Amp.	250	−9	—	—	9.5	8500	1900	16	10000	0.3	—
6S4	Triode	B.	9AC	6.3	0.6	—	—	—	Class-A$_1$ Amp.	250	−8	—	—	26	3600	4500	16	—	—	—
6T8	Triple-Diode Triode	B.	9E	6.3	0.45	1.5	1.1	2.4	Class-A$_2$ Amp.	250	−3	—	—	1.0	5800	1200	70	—	—	—
										100	−1	—	—	0.8	5400	1300	70	—	—	
12A4	Triode	B.	9AG	6.3 12.6	0.6 0.3	—	—	—	Class-A$_1$ Amp.	150	−17	—	—	30	1200	5200	6.5	—	—	—
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

TABLE XI — MINIATURE RECEIVING TUBES — *Continued*

Type	Name	Base	Socket Connections[1]	Fil. or Heater		Capacitance $\mu\mu$fd.			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
				Volts	Amp	In	Out	Plate-Grid												
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-A$_1$ Amp. Each Unit	250	– 2	—	—	10	10000	5500	55	—	—	—
				12.6	0.15	2.5[8]	0.35[8]	1.45[8]		180	– 1	—	—	11	9400	6600	62	—	—	
12AU6	Sharp Cut-off Pentode	B.	7CC	12.6	0.15	5.5	5.0	.0035	Class-A$_1$ 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-A$_1$ 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-A$_1$ 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-A$_1$ 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-A$_1$ 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*	Plate 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
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
12BH7	Dual Triode	B.	9A	6.3 12.6	0.6 0.3	3	2.6	2.4	Class-A$_1$ Amp.	250	– 9.5	—	—	11.5	—	3250	18	—	—	6SN7GT
12BK6	Duodiode Triode	B.	7BT	12.6	0.15	—	—	—	Class-A$_1$ Amp.	250	– 2	—	—	1.2	63000	1600	100	—	—	—
12BN6	Gated-beam Disc.	B.	7DF	12.6	0.15	4.2	3.3	.004	FM Disc.	Same as 6BN6										—
12BT6	Duodiode Triode	B.	7BT	12.6	0.15	—	—	—	Class-A$_1$ Amp.	Same as 6BT6										—
12BU6	Duodiode Triode	B.	7BT	12.6	0.15	—	—	—	Class-A$_1$ Amp.	Same as 6BU6										—
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	—	—	—	—
26BK6	Duodiode Triode	B.	7BT	26.5	0.07	—	—	—	Class-A$_1$ Amp.	Same as 6BK6										—
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	—	—	—	—
5670	Dual Triode	B.	8CJ	6.3	0.35	2.2	1.0	1.3	Class-A$_1$ Amp.	150	240*	—	—	8.2	—	5500	35	—	—	7F8
5686	Power Pentode	B	Fig. 29	6.3	0.35	6.4	4.0	0.11	Class-A$_1$ Amp.	250	–12.5	250	5	27	—	3100	—	9000	2.7	—
5687	Dual Triode	B.	9H	12.6	0.45	4	0.45	3.1	Class-A Amp.	250	–12.5	—	—	16	4000	4100	16.5	—	—	—
				6.3	0.9					120	– 2	—	—	34	2000	10000	20	—	—	
5722	Noise Generating Diode	B.	5CB	2/5.5	1.6	—	1.5	—	Noise Generator	200	—	—	—	35	—	—	—	—	—	—
5725	Semi remote Cut-off Pentode	B.	7CM	6.3	.175	—	—	—	Class-A$_1$ Amp.	120	– 2	120	3.5	5.2	—	3200	—	—	—	—
5726	Twin Diode	B.	6BT	6.3	0.3	—	3.2	—	Rectifier		Maximum a.c. voltage per plate = 117; Maximum d.c. Ma. per plate = 9.									—
5749	Remote Cut-off Pentode	B.	7BK	6.3	0.30	5.5	5.0	.0035	Class-A$_1$ Amp.	250	68*	100	4.2	11	1 Meg.	4400	—	—	—	—
5750	Pentagrid Converter	B.	7CH	6.3	0.30	Osc. Grid 20000Ω			Converter	250	– 1.5	100	7.5	2.6	1 Meg.	475	0.5[10]	—	—	—
5751	Dual Triode	B.	9A	12.6	.175	—	—	—	Class-A$_1$ Amp.	250	– 3	—	—	1.1	58000	1200	70	—	—	12SL7GT
5812	Beam Pentode	B.	7CQ	6.3	0.65	9	7.4	0.2	Class-A$_1$ Amp.	250	–23	250	1.8	40	55000	4100	—	—	—	—

TABLE XI—MINIATURE RECEIVING TUBES—*Continued*

Type	Name	Base	Socket Connections	Fil. or Heater		Capacitance $\mu\mu$fd.			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												
5814	Dual Triode	B.	9A	6.3 12.6	0.35 .175	1.6	0.5	1.5	Class-A$_1$ Amp.	250	− 8.5	—	—	10.5	6250	2200	19.5	—	—	12SN7GT
5079	Sharp Cut-off Pentode	B.	9AD	6.3	0.15	2.7	2.4	0.11	Class-A$_1$ Amp.	250	− 3	100	0.4	1.8	2 Meg.	1000	—	—	—	—
5915	Dual Control Sharp Cut-off Heptode	B.	7CH	6.3	0.3	7.2	8.6	0.3	Switch	30	− 5.5	75	8.25	6	—	—	—	—	—	—
5963	Dual Triode	B.	9A	12.6 6.3	0.15 0.3	1.9	—	1.5	Class-A$_1$ Amp.	67.5	0	—	—	7[1]	7850	2800	22	—	—	—
5964	Dual Triode	B.	7BF	6.3	0.45	2.1	—	1.3	Class-A$_1$ Amp.	100	50*	—	—	9.5[1]	6500	6000	39	—	—	—
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] Unless otherwise noted.
[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.
[10] Oscillator grid current Ma.

TABLE XII—SUB-MINIATURE TUBES

Type	Name	Base	Socket Connections	Fil. or Heater		Capacitance $\mu\mu$fd.			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												
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 Converter	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]	[2]	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]		1.25	0.03	—	—	—	Detector Amp.	22.5	0	22.5	0.12	0.35	—	—	—	—	—	2E41
2E42	Diode Pentode	[1]	[2]	1.25	0.03				Detector Amp.	22.5	0	22.5	0.12	0.35	250000	375	—	1 meg.	—	2E42
2G21	Triode Heptode	[1]	[2]	1.25	0.05	—	—	—	Converter	22.5	—	22.5	0.2	0.3	—	75	—	—		2G21
2G22	Converter		[2]	1.25	0.05				Converter	22.5	0	22.5	0.3	0.2	500000	60	—	—	—	2G22
6K4	Triode		[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		[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	0	30	0.06	0.3	1000000	325	—	—	—	CK501
										45	− 1.25	45	0.055	0.28	1500000	300				
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	0	30	0.07	0.17	1100000	140	—	—	—	CK505
										45	− 1.25	45	0.08	0.2	2000000	150				
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

TABLE XII—SUB-MINIATURE TUBES—*Continued*

Type	Name	Base	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												
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 μa	60 μa	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	Pentode 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	—	—	—	Radio 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
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

TABLE XII—SUB-MINIATURE TUBES—*Continued*

Type	Name	Base	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												
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-A₁ Amp.	100	560*	—	—	5.0	7400	2700	20	—	—	SN957A 5645
SN1006	Triode	[1]	[2]	6.3	0.15				Class-A₁ 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
5639	Video Pentode	[1]	8DL	6.3	0.45	9.5	7.5	0.10	Class-A₁ Amp.	150	100*	100	4.0	21	50K	9000	—	9000	1.0	5639
5641	Single Diode	[1]	6CJ	6.3	0.45	—	—	—	H. W. Rectifier				235 volts a.c. max.; 45 Ma. d.c. output.							5641
5643	Tetrode Thyratron	[1]	8DD	6.3	0.15	1.7	1.6	0.1	Relay Tube Grid Contr. Rect.	Peak anode volts=500; Inv. volts=500; Peak I_k=100 Ma.; Avg.=22 Ma.										5643
5644	Cold Cathode Diode	[1]	4CN	—	—	—	—	—	Voltage Reg.	Starting voltage=125 max. d.c. Operating voltage=95. Operating current=5–25 Ma. Regulation=4 volts approx.										5644
5647	Single Diode	[1]	81	6.3	0.15	2.2	—	—	H. W. Rectifier			150 volts a.c. max; 9 Ma. d.c. output.								5647
5718	U.h.f. Medium-Mu Triode	[1]	8DK	6.3	0.15	2.2	0.7	1.4	Class-A₁ Amp.	150	180*	—	—	13	4150	6500	27	—	—	5718
									U.h.f. Oscillator	150	−12	Fmc.=500		20		Ig=3.7 Ma.			0.9	
5719	Hi-Mu Triode	[1]	8DK	6.3	0.15	2.4	0.6	0.7	Class-A₁ Amp.	150	680*	—	—	1.7	26000	2700	70	—	—	5719
5840	U.h.f. Sharp Cut-off Pent.	[1]	8DL	6.3	0.15	4.2	4.0	0.015	Class-A₁ Amp.	100	150*	100	2.4	7.5	230K	5000	—	—	—	5840
5896	U.h.f. Dual Diode	[1]	8DJ	6.3	0.3	3.0	—	—	Det.-Rectifier	150 volts a.c. max.; 9 Ma. d.c. output per plate.										5896
5897	U.h.f. Medium-Mu Triode	[1]	8DK	6.3	0.15	2.2	0.7	1.4	Class-A₁ Amp.	150	180*	—	—	13	4150	6500	27	—	—	5897
									U.h.f. Oscillator	150	−12	—	—	20	Ig=3.7 Ma. Fmc.=500.				0.9	
5898	Hi-Mu Triode	[1]	8DK	6.3	0.15	2.4	0.6	0.7	Class-A₁ Amp.	150	680*	—	—	1.7	26000	2700	70	—	—	5898
5899	U.h.f. Semi-Remote Pent.	[1]	8DL	6.3	0.15	4.4	4.0	0.015	Class-A₁ Amp.	100	120*	100	2.2	7.2	260K	4500	—	—	—	5899
5900	U.h.f. Semi-Remote Pent.	[1]	8DL	6.3	0.15	4.4	4.0	0.015	Class-A₁ Amp.	100	120*	100	2.2	7.2	260K	4500	—	—	—	5900
5901	U.h.f. Sharp Cut-off Pent.	[1]	8DL	6.3	0.15	4.2	4.0	0.015	Class-A₁ Amp.	100	150*	100	2.4	7.5	230K	5000	—	—	—	5901
5902	Audio Beam Pentode	[1]	8DL	6.3	0.15	6.5	7.5	0.11	Class-A₁ Amp.	110	270*	110	2.2	30	15K	4200	—	3000	1.0	5902
5903	U.h.f. Dual Diode	[1]	8DJ	26.5	0.075	3.0	—	—	Det.-Rectifier	150 volts a.c. max.; 9 Ma. d.c. output per plate.										5903
5904	U.h.f. Medium-Mu Triode	[1]	8DK	26.5	0.045	2.2	0.8	1.8	Class-A₁ Amp.	26.5	−3.5	—	—	3	3800	5000	19	—	—	5904
									U.h.f. Oscillator	26.5	0	—	—	20	Ig=7.5 Ma. Fmc.=400.				0.06	
5905	U.h.f. Sharp Cut-off Pent.	[1]	8DL	26.5	0.045	4.4	4.2	0.015	Class-A₁ Amp.	26.5	2.2[6]	26.5	0.9	2.3	110K	2850	—	—	—	5905
5906	U.h.f. Sharp Cut-off Pent.	[1]	8DL	26.5	0.045	4.2	4.0	0.015	Class-A₁ Amp.	100	150*	100	2.4	7.5	230K	5000	—	—	—	5906
5907	U.h.f. Remote Cut-off Pent.	[1]	8DL	26.5	0.045	4.4	4.0	0.015	Class-A₁ Amp.	26.5	2.2[6]	26.5	1.1	2.7	125K	3000	—	—	—	5907
5908	U.h.f. Pentode	[1]	8DC	26.5	0.045	4.4	4.6	0.08	Class-A₁ Amp.	26.5	2.2[6]	26.5	1.6	2.3	30K	1750	—	—	—	5908
									Mixer	26.5	2.2[6]	26.5	1.6	1.0	100K	800	—	—	—	
5916	U.h.f. Pentode		8DC	26.5	0.045	4.2	4.0	0.015	Class-A₁ Amp.	100	150*	100	3.4	4.4	130K	3000	—	—	—	5916
									Mixer	100	150*	100	4.6	2.5	400K	1100	—	—	—	

* 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. [6] Grid leak resistor, megohms.

TABLE XIII—CONTROL AND REGULATOR TUBES

Type	Name	Base	Socket Connections	Cathode	Fil. or Heater		Use	Peak Anode Voltage	Max. Anode Ma.	Minimum Supply Voltage	Operating Voltage	Operating Ma.	Grid Resistor	Tube Voltage Drop	Type
					Volts	Amp.									
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

TABLE XIII—CONTROL AND REGULATOR TUBES—*Continued*

Type	Name	Base	Socket Connections	Cathode	Fil. or Heater		Use	Peak Anode Voltage	Max. Anode Ma.	Minimum Supply Voltage	Operating Voltage	Operating Ma.	Grid Resistor	Tube Voltage Drop	Type
					Volts	Amp.									
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[6]	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							885
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
5663	Tetrode Thyratron	7-pin B	7CE	Htr.	6.3	0.15	Control and Relay	Max. peak inv. volts = 500; Peak Ma. = 100; Avg. Ma. = 20.							5663
5823	Gas Triode	7-pin B.	4CK	Cold	—	—	Relay or Trigger	Max. peak inv. volts = 200; Peak Ma. = 100; Avg. Ma. = 25.							5823
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[7]	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.

#Discontinued.

[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 Connections	Heater		Use	Size	Anode No. 2 Voltage	Anode No. 1 Voltage	Cut-Off Grid Voltage	Grid No. 2 Voltage	Ion-Trap Ma.	Max. Input Voltage[1]	Focus Coil Ma.	Deflection Sensitivity[6]		Anode No. 3 Voltage	Pattern Color	Type
			Volts	Amp.										D_1 D_2	D_3 D_4			
2AP1[7]-11	Electrostatic Cathode-Ray	11B	6.3	0.6	Oscillagraph 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]	—		

TABLE XIV—CATHODE-RAY TUBES AND KINESCOPES—*Continued*

Type	Name	Socket Connections	Heater		Use	Size	Anode No. 2 Voltage	Anode No. 1 Voltage	Cut-Off Grid Voltage	Grid No. 2 Voltage	Ion-Trap Ma.	Max. Input Voltage [1]	Focus Coil Ma.	Deflection Sensitivity [8]		Anode No. 3 Voltage	Pattern Color	Type
			Volts	Amp.										$D_1 D_2$	$D_3 D_4$			
3AP1/906-P1-4-5-11[7]	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	3"	1500	430	−50	—	—	550	—	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			
30P1	Electrostatic Cathode-Ray	Fig. 49	6.3	0.6	Oscillograph	3"	2000	575	−60	—	—	550	—	200[3]	148[3]	—	Green	30P1
							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]			
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	—	—	—	—	146/198[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	—	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	—	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 Yellow 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

TABLE XIV—CATHODE-RAY TUBES AND KINESCOPES—*Continued*

Type	Name	Socket Connections	Heater		Use	Size	Anode No. 2 Voltage	Anode No. 1 Voltage	Cut-Off Grid Voltage	Grid No. 2 Voltage	Ion-Trap Ma.	Max. Input Voltage [1]	Focus Coil Ma.	Deflection Sensitivity [6]		Anode No. 3 Voltage	Pattern Color	Type
			Volts	Amps.										D_1 D	D_3 D_4			
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	—	—	—	—	—	—	—	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	—	—	—		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
7MP7	Electromagnetic Cathode Ray	12D	6.3	0.6	Oscillograph Radar	7″	—	7000	-27/-63	250	—	—	85	—	—	—	G'rnish-Yellow	7MP7
							—	4000	-27/-63	250	—	—	62	—	—	—		
8AP4	Electromagnetic Picture Tube	12H	6.3	0.6	Television	8″	—	7000	-27/-63	—	45[8]	—	115	—	—	—	White	8AP4
8BP4	Electrostatic Picture Tube	14G	6.3	0.6	Television	8″	—	2400	-72/-168	6000	—	—	—	146/198[3]	124/168[3]	—	White	8BP4
							—	1620	-72/-168	6000	—	—	—			—		
9AP4/ 1804-P4	Electromagnetic Kinescope	6AL	2.5	2.1	Television	9″	7000	1425	− 40	250	—	—	—	—	—	—	White	9AP4/ 1804-P4
							6000	1225	− 38									
9CP4	Electromagnetic Kinescope	4AF	2.5	2.1	Television	9″	7000	—	−110	—	—	—	—	—	—	—	White	9CP4
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	—	—	—	—	—	—	White	10BP4
10EP4	Magnetic-Focus Cathode-Ray	12D	6.3	0.6	Television	10½″	—	8000	− 45	250	—	—	—	—	—	—	White	10EP4
10FP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	10″	—	9000	-27/-63	250	—	—	—	—	—	—	White	10FP4
10HP4	Electrostatic Cathode-Ray	14G	6.3	0.6	Television	10″	—	5000	-60/-140	1800	—	—	—	130[3]	100[3]	—	White	10HP4
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-A	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	12″	—	11000	-27/-63	250	—	—	—	—	—	—	White	12KP4-A
12LP4	Electromagnetic Kinescope	12D	6.3	0.6	Television	12″	—	11000	-27/—63	250	—	—	—	—	—	—	White	12LP4
14BP4	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	14″	—	11000	-27/-63	250	120	—	110	—	—	—	White	14BP4
15AP4	Electromagnetic Cathode-Ray	12D	6.3	0.6	Television	15″	—	8000	− 45	250	—	—	—	—	—	—	White	15AP4
15CP4	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	15″	—	9000	− 45	250	109	—	115	—	—	—	White	15CP4
15DP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	15″	—	13000	-27/-63	250	105	—	146	—	—	—	White	15DP4
16AP4	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	16″	—	12000	-33/-77	300	—	—	—	—	—	—	White	16AP4
16EP4A	Electromagnetic Picture Tube	12D	6.3	0.6	Television	16″	—	12000	-33/-77	300	—	—	105	—	—	—	White	16EP4A
16FP4	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	16″	—	13000	-27/-63	250	105	—	146	—	—	—	White	16FP4
16GP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	16″	—	12000	-33/-77	300	23[8]	—	100	—	—	—	White	16GP4
16HP4	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	16″	—	12000	-33/-77	300	120	—	110	—	—	—	White	16HP4
16JP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	16″	—	11000	-27/-63	250	120	—	115	—	—	—	White	16JP4
16KP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	16″	—	14000	-33/-77	300	30[8]	—	90	—	—	—	White	16KP4
16LP4	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	16″	—	12000	-33/-77	300	120	—	110	—	—	—	White	16LP4
16RP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	16″	—	12000	-33/-77	300	120	—	100	—	—	—	White	16RP4
16TP4	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	16″	—	12000	-33/-77	300	45[8]	—	115	—	—	—	White	16TP4
16UP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	16″	—	12000	-27/-63	300	23[8]	—	100	—	—	—	White	16UP4
19AP4	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	16″	—	13000	-27/-63	250	105	—	146	—	—	—	White	19AP4

TABLE XIV—CATHODE-RAY TUBES AND KINESCOPES—*Continued*

Type	Name	Socket Connections	Heater		Use	Size	Anode No. 2 Voltage	Anode No. 1 Voltage	Cut-Off Grid Voltage	Grid No. 2 Voltage	Ion-Trap Ma.	Max. Input Voltage[1]	Focus Coil Ma.	Deflection Sensitivity[6]		Anode No. 3 Voltage	Pattern Color	Type
			Volts	Amps.										D_1 D	D_3 D_4			
19AP4A	Electromagnetic Picture Tube	12D	6.3	0.6	Television	16″	—	12000	-33/-77	300	75	—	140	—	—	—	White	19AP4A
19FP4	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	16″	—	13000	-27/-68	250	100	—	100/130	—	—	—	White	19FP4
20BP4	Electromagnetic Cathode-Ray	12D	6.3	0.6	Television	20″	—	15000	– 45	250	—	—	—	—	—	—	White	20BP4
22AP4	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	22″	—	14000	-33/-77	300	35[8]	—	117	—	—	—	White	22AP4
902[7]	Electrostatic Cathode-Ray	Fig. 1	6.3	0.6	Oscillograph	2″	600	150	– 60	—	—	350	—	0.19	0.22	—	Green	902
903[5]	Electromagnetic Cathode-Ray	6AL	2.5	2.1	Oscillograph	9″	7000	1360	–120	250	—	—	—	—	—	—	Green	903
904	Electrostatic-Magnetic Cathode-Ray	Fig. 3	2.5	2.1	Oscillograph	5″	4600	970	– 75	250	—	4000	—	0.09	—	—	Green	904
905[7]	Electrostatic Cathode-Ray	Fig. 6	2.5	2.1	Oscillograph	5″	2000	450	– 35	—	—	1000	—	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	—	0.041	0.051	—	Green	912
913	Electrostatic Cathode-Ray	Fig. 1	6.3	0.6	Oscillograph	1″	500	100	– 65	—	—	250	—	0.07	0.10	—	Green	913
914[7]	Electrostatic Cathode-Ray	Fig. 12	2.5	2.1	Oscillograph	9″	7000	1450	– 50	250	—	3000	—	0.073	0.093	—	Green	914
1800[5]	Electromagnetic Kinescope	6AL	2.5	2.1	Television	9″	6000	1250	– 75	250	—	—	—	—	—	—	Yellow	1800
1801[5]	Electromagnetic Kinescope	Fig. 13	2.5	2.1	Television	5″	3000	450	– 35	—	—	—	—	—	—	—	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	Electrastatic Cathode-Ray	Fig. 1[4]	2.5	2.1	Television	5″	2000	1000	– 35	200	—	—	—	0.5	0.56	—	—	2005
24-XH	Electrostotic Cathode-Ray	Fig. 1	6.3	0.6	Oscilloscope	2″	600	120	– 60	—	—	—	—	0.14	0.16	—	Blue	24-XH

[1] Between Anode No. 2 and any deflecting plate.

[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."

[8] Ion-trap gausses.

TABLE XV—RECTIFIERS—RECEIVING AND TRANSMITTING

See also Table XIII—Control and Regulator Tubes

Type No.	Name	Base	Socket Connections	Cathode	Fil. or Heater		Max. A.C. Voltage Per Plate	D.C. Output Current Ma.	Max. Inverse Peak Voltage	Peak Plate Current Ma.	Type
					Volts	Amp.					
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
1V2	Half-Wave Rectifier	9-pin B.	9U	Fil.	.625	0.3	—	0.5	7500	10	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.	9Y	Fil.	1.25	0.2	—	1	15000	10	HV
1X2A	Half-Wave Rectifier	9-pin B.	9Y	Fil.	1.25	0.2	—	1.1	20000	11	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.	4P	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.	5Q	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
6AX5GT	Full-Wave Rectifier	6-pin O.	6S	Htr.	6.3	1.2	450	125	1250	375	HV
6AX6G	Full-Wave Rectifier	7-pin O.	7Q	Htr.	6.3	2.5	350	250	1250	600	HV
6U4GT	Half-Wave Rectifier	5-pin O.	4CG	Htr.	6.3	1.2	—	138	1375	660	HV
6V4	Full-Wave Rectifier	9-pin B.	9M	Htr.	6.3	0.6	350	90	—	—	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
25W4GT	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

TABLE XV—RECTIFIERS—RECEIVING AND TRANSMITTING—*Continued*

See also Table XIII—Control and Regulator Tubes

Type No.	Name	Base	Socket Connections	Cathode	Fil. or Heater		Max. A.C Voltage Per Plate	D.C. Output Current Ma.	Max. Inverse Peak Voltage	Peak Plate Current Ma.	Type
					Volts	Amp.					
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[6]	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
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
50AX6G	Full-Wave Rectifier	7-pin O.	7Q	Htr.	50	0.3	350	250	1250	600	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[6]	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 5.	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

TABLE XV—RECTIFIERS—RECEIVING AND TRANSMITTING—*Continued*

See also Table XIII—Control and Regulator Tubes

Type No.	Name	Base	Socket Connections	Cathode	Fil. or Heater		Max. A.C. Voltage Per Plate	D.C. Output Current Ma.	Max. Inverse Peak Voltage	Peak Plate Current Ma.	Type
					Volts	Amp.					
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 Rectifier	4-pin M.	4P	Fil.	5.0	5.5	10000	100	40000	750	HV
					5.8	6.5	12500	100	40000	750	
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	Cathade		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	1.0	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**	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	180	—	—	—	—	—	—
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-BAG	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
2C36	5	6.3	0.4	1500[5]	—	—	25	1.4	2.4	0.36	1200	N.	Fig. 36	Plate-Pulsed 1000-Mc. Osc.	1000[5]	0	900[5]	—	—	—	200[5]
2C37 5766 5767	5	6.3	0.4	350	—	—	25	1.4	1.85	0.02	3300	N.	Fig. 36	1000-Mc. C.W. Oscillator	150	3000 **	15	3.6	—	—	0.5
5764	5	6.3	0.4	1500[5]	11.5	—	25	1.4	1.85	0.02	3300	N.	Fig. 36	Plate-Pulsed 3300-Mc. Osc.	1000[5]	0	1300[5]	—	—	—	200[5]
5765	5	6.3	0.4	350	—	—	25	1.3	2.1	0.03	2900	N.	Fig. 36	1900-Mc. C.W. Oscillator	180	10000 **	25	—	—	—	0.225
5675	5	6.3	0.135	165	30	8	20	2.3	1.3	0.09	3000	N.	Fig. 36	Grounded-Grid Osc.	120	— 8	25	4	—	—	0.05
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
5876	6.25	6.3	0.135	300	25	—	56	2.5	1.4	0.035	1700	N.	Fig. 36	Grounded-Grid Oscillator	250	— 2	23	3	—	—	0.75
														Frequency Multiplier	300	— 70	17.3	7	—	—	2.0
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

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.											
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
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[1]	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[1] 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	Class-C Amp. (Telegraphy)	Characteristics similar to 25T						
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

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.											
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						
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 VT-191	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

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.											
HY40[1]	40	7.5	2.25	1000	125	25	25	6.1	5.6	1.0	60	M.	3G	Class-C Amp. (Tolegraphy)	1000	− 90	125	20	5.0	—	94
														Class-C Amp. (Telephony)	850	− 90	125	25	5.0	—	82
														Grid-Modulated Amp.	1000	—	125	—	—	—	20
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 Amo. 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	Closs-C Amp. (Telegraphy)	850	− 48	110	15	2.5	—	70
														Class-C Amp. (Telephany)	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 3-50D4 35TG	50	5.0	4.0	2000	150	50	39	4.1 2.5	1.8 1.8	0.3 0.4	100 100	M. M.	3G 2D	Class-C Amp. (Telegraphy)	2000	−135	125	45	13	—	200
														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)	15	−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		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.											
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
8415W	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	22
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	1500	−4.5	350 [8]	88 [8]	6.5 [8]	10500	400

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.											
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	0.3	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. (Telephany)	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		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.											
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	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	—	—	15.5	2.7	2.3	0.35	150	N.	T-4B	Class-C Amp. (Telegraphy)	2000	−340	210	67	25	—	315
														Class-B Amp. (Audio) [7]	1500	−125	242	44	7.3	3000	200
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		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.											
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. (Telegruphy)	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. (Audia)[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. (Telephany)	1000	−125	150	50	—	—	100
														Class-B Amp. (Audia)[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
2C39A	100	6.3	1.0	1000	80	50	100	6.5	1.95	.035	500	N.	—	Class-C Amplifier	800	− 45	80	45	6	—	27
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 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.											
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
AX9900/ 5866[12]	135	6.3	5.4	2500	200	40	25	5.8	5.5	0.1	150	N.	Fig. 5	Class-C Amp. (Telegraphy)	2500	—200	200	40	16	—	390
														Class-C Amp. (Telephony)	2000	—225	127	40	16	—	204
														Class-B (Audio)[7]	2500	— 90	80/330	350[9]	14[8]	15680	560
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	12.51 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.-Oscillatar	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. (Audia)[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	3.8	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		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.											
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	8.0	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	260	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 3-200A3	200	10	5.0	3500	250	50	25	3.6	3.3	0.29	150	N.	Fig. 52	Class-C Amp. (Telegraphy)	3000	–220	222	25	11	—	466
														Class-C Amp. (Telephony)	2500	–300	200	35	19	—	375
														Class-B (Audio)[7]	2000	– 50	120/500	520[9]	25[8]	8500	600
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 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.											
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.
** Grid resistor 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] Peak valves.
[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 Dissipation Watts	Cathode		Max. Plate Voltage	Max. Screen Voltage	Max. Screen Dissipation Watts	Interelectrode Capacitances (μμfd.)			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
		Volts	Amp.				Grid to Fil.	Grid to Plate	Plate to Fil.															
3A4	2.0	1.4 2.8	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	4.5	2.8 1.4	0.11 0.22	180	135	0.9	7.5	0.3	5.5	50	L.	6BB	Class-C Amp. (Telegraphy)	150	135	—	−20	23	6.0	1.0	—	0.25	—	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	15	4.0	3.0	—	—	—	4.0
5A6	5.0	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
5686	7.5	6.3	0.35	250	250	3.0	6.4	0.11	4.0	160	B.	Fig. 29	Class-C Amp. (Telegraphy)	250	250	—	−50	40	10.5	2.0	—	0.15	—	6.5
														250	180	—	−30	30	6.5	2.0	—	0.10	—	5.0
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	4.0	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
5812	10	6.0	0.65	300	250	2.5	9.0	0.2	7.4	165	B.	7CQ	Class-C Amp. (Telegraphy)	300	200	—	−45	55	3.0	0.75	—	1.5	—	7.0
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. (Telephany)	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	0.2	13	10	O.	7AC	Class-C Amp. (Telegraphy)	400	275	—	−100	50	11	5.0	—	—	—	14
							8.0	0.5	6.5				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
													Class-C Amp. (Telephony)	500	180	—	−50	54	9.0	2.5	35500	0.15	—	18
	9.0			500	200	2.3							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

TABLE XVII—TETRODE AND PENTODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissipation Watts	Cathode		Max. Plate Voltage	Max. Screen Voltage	Max. Screen Dissipation Watts	Interelectrode Capacitances (μμfd.)			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
		Volts	Amp.				Grid to Fil.	Grid to Plate	Plate to Fil.															
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
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
AX-9905[3]	16	6.3	0.68	400	250	5	8.5	0.05	3.3	186	O.	Fig. 34	Class-C Amplifier	400	250	—	–80	80	6	3.5	—	0.39	—	20.8
														250	175	—	–70	80	6.5	4.2	—	0.26	—	16.9
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	21	6.3	0.9	400	300	3.5	10	0.4	12	10	O.	7AC	Class-C Amp.-Oscillator	400	300	—	–125	100	12	5.0	—	—	—	28
6L6G							11.5	0.9	9.5				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
5881	23	6.3	0.9	400	300	3	—	—	—	—	O.	7AC	Class-C Amplifier	Same as 6L6										
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

TABLE XVII—TETRODE AND PENTODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissipation Watts	Cathode		Max. Plate Voltage	Max. Screen Voltage	Max. Screen Dissipation Watts	Interelectrode Capacitances ($\mu\mu$fd.)			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
		Volts	Amp.				Grid to Fil.	Grid to Plate	Plate to Fil.															
HY61	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
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 807W 5933 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[8]	—	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	1.0	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
AX-9903[3]	40	6.3 12.6	1.8 0.9	600	250	7	6.7	0.08	2.1	150	N.	Fig. 10	Class-C Amp. (Telegraphy)	600	250	—	−80	200	16	2	—	0.2	—	80
													Class-C Amp. (Telephony)	600	250	—	−100	200	24	8	—	1.2	—	85
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	18	—	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]	—	80
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]	40	12.6 6.3	1.125 2.25	750	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

TABLE XVII—TETRODE AND PENTODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode		Max. Plate Volt-age	Max. Screen Volt-age	Max. Screen Dissi-pation Watts	Interelectrode Capacitances (μμfd.)			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
		Volts	Amp.				Grid to Fil.	Grid to Plate	Plate to Fil.															
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
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	2.8	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 4D32	50	25.2 12.6 6.3	0.8 1.6 3.75	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
												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.	7BM	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

TABLE XVII—TETRODE AND PENTODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissipation Watts	Cathode		Max. Plate Voltage	Max. Screen Voltage	Max. Screen Dissipation Watts	Interelectrode Capacitances (μμfd.)			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
		Volts	Amp.				Grid to Fil.	Grid to Plate	Plate to Fil.															
HK257 HK257B	75	5.0	7.5	4000	750	25	13.8	0.04	6.7	75 120	J.	7BM	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	—	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
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-3B	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		9	—	3.3	—	300
													Class-AB$_2$ Amp. (Audio) [6]	2500	350	—	− 43	93/260	0/6	178 [8]	—	1 0	22200	400
4E27A/ 5-125B	125	5.0	7.5	4000	750	20	10.5	0.08	4.7	75	J.	7BM	Class-C Amp. (Telegraphy)	3000	500	60	−200	167	5	6	—	1.6	—	375
														1500	500	60	−130	200	11	8	—	1.6	—	215
														1000	750	0	−170	160	21	3	—	0.6	—	115
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
													Suppressar-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	2.0	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
4X-150G	150	2.5	6.25	1250	300	15	16.1	0.02	4.7	165	N.	—	Class-C Amp. (Telegraphy)	1250	250	—	− 90	200	20	11	—	1.2	—	195
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

TABLE XVII—TETRODE AND PENTODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode		Max. Plate Volt-age	Max. Screen Volt-age	Max. Screen Dissi-pation Watts	Interelectrode Capacitances (μμfd.)			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
		Volts	Amp.				Grid to Fil.	Grid to Plate	Plate to Fil.															
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		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
		Volts	Amp.										
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
2K28[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 Oscillotor	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 Oscillotor *	1000	60	75	+24	−510	60	—	0.75
2K42[3]	3300–4200	6.3	1.3	Fig. 59	Reflex Oscillotor *	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 Oscillotar *	1000	90	80	0	−300	70	—	1–2

TABLE XVIII—KLYSTRONS—*Continued*

Type	Freq. Range-Mc.	Cathode		Base Connections	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
		Volts	Amp.										
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
6BL6	1250–6000	—	—	—	Reflex Oscillator	350	—	—	+ 1	0/−400	25	—	—
6BM6	550–3000	—	—	—	Reflex Oscillator	350	—	—	+ 1	0/−600	20	—	—
707B [5]	1200–3750	6.3	0.65	Fig. 61	Reflex Oscillator	300 [7]	45	—	300	−155/−290	30	—	0.140
5D1103	1250–6000	—	—	—	Reflex Oscillator	350	—	—	+10	0/−400	25	—	—
5D1104	550–3000	—	—	—	Reflex Oscillator	350	—	—	+10	0/−600	22	—	—
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—CRYSTAL TRIODES

Type	Maximum Ratings				Typical Operation								
	Collector			Emitter	Collector			Emitter			Transconductance μ-Mhos.	Power Gain Db.	Power Output M. Watts
	Volts	Ma.	Dissipation M. Watts	Ma.	Volts	Ma.	Z Ohms	Volts	Ma.	Z Ohms			
CK703	−70	4	200	10	−30	2	10000	0.2	0.75	500	5000	16	2

TABLE XX—CAVITY MAGNETRONS

Type	Class	Band or Range Mc.	Heater		Maximum Ratings					Typical Operation				Peak Pwr. Output KW.
			Volts	Amps.	Anode KV.	Anode Amps.	Duty Cycle	Input Watts	Anode KV.	Anode Amps.	Field Gauss	Pulse μ Sec.	P.P.S.	
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
2J42	1	9345–9405	6.3	0.5	5.7	6.5	.001	—	—	—	4800	2.5	—	14
2J42A	1	9345–9405	6.3	0.5	8.0	7.0	.001	—	—	—	6500	2.5	—	35
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
3J31	1	23744–24224	6.0	1.9	15.0	14.0	.0005	—	—	—	7600	1.0	—	54
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
4J50	1	9345–9405	13.6	3.5	23.0	27.5	.004	—	—	—	6300	0.5	—	300
4J52	1	9345–9405	12.6	1.9	16.0	15.0	.002	—	—	—	5000	6.0	—	120
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
4J78	1	9003–9168	13.6	3.5	23.0	27.5	.004	—	—	—	6300	0.5	—	300
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.

The Catalog Section

☆ ☆ ☆

In the following pages is a catalog-file of the principal manufacturers and the principal distributors who serve the radio communication and industrial fields. Appearance in these pages is by invitation—space has been sold only to those dependable firms whose established integrity and whose products have earned the approval of the American Radio Relay League.

INDEX OF ADVERTISERS

★ CATALOG SECTION ★

The Radio Amateur's Handbook

Model TV-2030

Features life-size 20" rectangular, dark-screen picture tube, 10" high-fidelity speaker and built-in "Televane" antenna in a handsome genuine blond mahogany veneer console cabinet. Dark mahogany finish (Model TV-2029) also available.

National

NC

COMPONENTS TELEVISION
COMMUNICATION RECEIVERS

National Television is engineered and built with the same superb skill and custom craftsmanship that have made National communication equipment known and respected throughout the world. The latest turret-type tuners, a full 4 mcs. video bandwidth and dark-screen rectangular picture tubes are just a few of the many fine features that distinguish the 1951 line of ten 17" and 20" models. See your National dealer or write for TV literature.

The superb National TV chassis engineered for use with 16", 17", 19" or 20" picture tubes. Latest turret-type tuner, double-tuned RF and 4 mcs. video bandwidth for pictures of unsurpassed strength and clarity. Three-unit design provides flexible mounting for custom installations.

HRO-50

$359*

(including coils A, B, C, D. Less speaker.)

Latest and Finest Version of the World-Famous HRO Series!

Compare the characteristics and features of the new HRO-50 and see why, once again, the HRO sets the standard of receiver performance! You'll appreciate the convenience of the new HRO-50, too — the new edge-lighted, direct reading dial and the insulated, heavy-duty, built-in power supply section. For thrilling performance, be sure to see and try the new HRO-50!

COVERAGE: 50-430 kc., 480 kc.-35 mc. Voice, CW, NFM (with adaptor).

FEATURES: Sensitivity of 1 mv. or better at 6 db. sig./noise. Selectivity variable from 13 kc. overall to app. 1200 cps. at 40 db. Negligible drift after warm-up. Micrometer dial for logging. Provision for crystal calibrator unit. Variable ant. trimmer. Lively S-meter. Min. tubes in front end and high freq. osc. Osc. circuits not disabled when receiver in send position. High-fidelity push-pull audio with phono jack. BFO switch separated from BFO freq. control. Dimmer illumination control. Accessory socket for Select-O-Ject.

CONTROLS: Bandswitch, Oscillator, Tone, Ant. Trimmer, Dimmer, AVC, Limiter, AF Gain, Calibration, CWO, Phasing, Selectivity, On-Off, RF Gain, AM-NFM-PHONO.

TUBE COMPLEMENT: 6BA6, 1st r.f.; 6BA6 2nd r.f.; 6BE6, mixer; 6C4 h.f. oscillator; 6K7, 1st i.f.; 6K7 2nd i.f.; 6H6 det. & a.v.c.; 6H6, a.n.l.; 6SJ7, 1st audio; 6SN7, phase splitter and S-meter amp.; 6V6GT (2) p.p. audio; 5V4G, rect.; 6J7, b.f.o.; OB2, volt. reg. Accessories: Crystal Calibrator, 6AK6; NFM Adaptor, 6SK7, i.f. amp., 6H6, ratio det.; Select-O-Ject, 12AT7 (2).

HRO-50C
(HRO-50 receiver with rack, speaker and 10-coil compartment. Coils A, B, C, D included.)

SIZE: Table 19¾" wide x 10⅛" high x 16½" deep. Rack: 19" wide x 10½" high x 17$\frac{7}{16}$" from rear of front panel incl. 1⅛" handle.

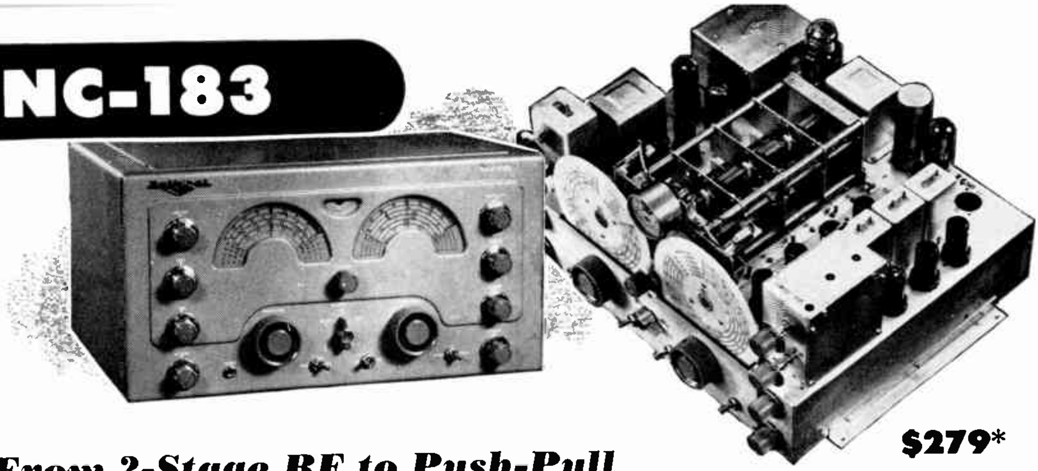
Write for list of accessories.

National Company, Inc.

*Slightly higher west of the Rockies.

NATIONAL COMPANY, Inc.

NC-183



$279*
(less speaker)

From 2-Stage RF to Push-Pull Audio . . . Designed to Out-Perform!

COVERAGE: Continuous from 540 kcs. to 31 mcs. plus 48 to 56 mcs. for 6-meter reception.

FEATURES: Two tuned R.F. stages. Voltage regulated osc. and BFO. Main tuning 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. New-type noise limiter. High fidelity push-pull audio. Accessory socket for NFM adaptor or other unit, such as crystal calibrator.

CONTROLS: CWO Switch, CWO pitch, Tone, AF Gain, Main Tuning, Bandspread, Ant. Trimmer, Bandswitch, Send-Receive, Phono-Radio, Selectivity, Phasing, Limiter, RF Gain.

TUBE COMPLEMENT: 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.; 1-5U4G rect.

NC-125

$149.50*
(less speaker)

Moderate-Priced Receiver with Built-In Selecto-O-Ject

COVERAGE: 560 kcs. to 35 mc. in 4 bands. Voice or CW.
FEATURES: Edge-lighted direct-reading scale with amateur, police, foreign, ship frequencies clearly marked. Sensational National Select-O-Ject built-in. Exceptional sensitivity on all bands. Lively S-meter reads S9 to 50 mv. signal. AVC, ANL, jack for phono or NFM adaptor, volt. reg., stabilized osc., audio essentially flat to 10,000 c.p.s.

CONTROLS: Main Tuning, Bandspread, Freq. (SOJ), Boost (SOJ), Send-Receive, Pitch, CWO-MVC-AVC-ANL, AF Gain, Tone, Trimmer, Bandswitch, RF Gain.

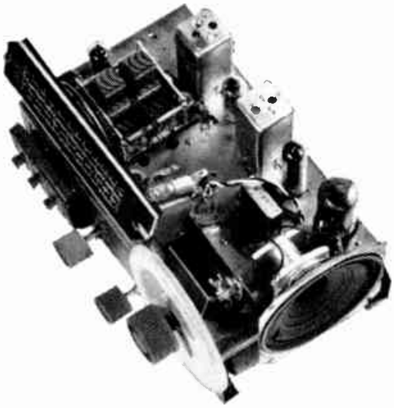
TUBE COMPLEMENT: 6SG7 RF amp, 6SB7-Y osc.-mixer, 6SG7 1st IF, 6SG7 2nd IF, 6H6 2nd det-AVC-ANL, 6SL7GT phase shifter, 6SL7GT boost-reject aud. amp., 6SL7GT 1st aud.-CWO, 6V6GT aud. output, OD3/VR-150 volt. reg., 5Y3GT rect.

**Slightly higher west of the Rockies.*

61 SHERMAN ST., MALDEN, MASS.

SW-54

Sleek Low-Priced Beauty is Most Compact General Coverage Receiver Ever Built!



USES MINIATURE TUBES AND UNIQUE BANDSPREAD DIAL

New miniature tubes make possible new sensitivity and performance. Unique plastic bandspread dial is adjustable to assure complete logging accuracy.

Here is National's latest engineering triumph! A complete superheterodyne receiver covering all major broadcast and shortwave bands that is smaller than the average table radio! New design makes possible a standard of performance never before achieved in so compact a receiver!

COVERAGE: Entire frequency range from 540 kc. to 30 mc. Voice, music or code.

FEATURES: Sensitive and selective superhet circuit. Slide rule general coverage dial with police, foreign, amateur and ship bands clearly marked. Separate bandspread and logging scale usable over entire range.

CONTROLS: Main tuning, Bandspread, On-Off and Volume, Receive-Standby, Bandswitch, AM-CW, Speaker-Phones.

TUBE COMPLEMENT: 12BE6, converter; 12BA6, CW osc. — IF amp.; 12AV6, 2nd det.-1st aud. — A.V.C.; 50C5, audio output; 35Z5, rectifier.

SIZE: 11" wide, 7" high, 7" deep.

(Complete with built- in speaker and power supply)

NATIONAL COMPANY, Inc.

NC-57B

Popular Low-Priced Receiver that Has Set New Performance Records!

NC-57C Same as NC-57B except that frequency range is 190-410 kcs. and 540 kcs. 35 mcs. **$99.50***

NC-57M Same coverage as NC-57C but wired for AC/DC operation. **$99.50***

(with built-in speaker & power supply)

COVERAGE: Phone and CW from 550 kcs. to 55 mcs. in 5 bands. **FEATURES:** Volt. stab. osc. keeps signal steady regardless of line voltage changes, aut. noise lim., provision for connecting S-meter and other accessories, ant. terminals for single, double or co-ax ant. lead. **CONTROLS:** Main Tuning, Bandspread, Band Switch, RF Gain, RF Trimmer, BFO-MVC-AVC, ANL Switch, AF Gain, BFO Pitch, Tone Control and On-Off Switch. **TUBE COMPLEMENT:** 6SG7 RF amp., 6SB7Y conv., 2-6SG7 IF amp., 6H6 Det., AVC, ANL, 6SL7GT Audio amp., BFO; 6V6GT Audio amp., 5Y3GT rect., VR-150 voltage rect. **$99.50***

SELECT-O-JECT

Boosts or Rejects Any Selected Frequency 38db!

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.! **$24.95***

TV BOOSTER

Makes the Difference Between A Picture and No Picture!

Adds a stage of RF amplification to average TV set. If signal is low, but perceptible, this booster will aid materially in increasing brightness and definition. Utilizes turret tuner for exceptionally high gain and uniform bandwidth on all channels. Housed in smart metal cabinet finished in special wear-resistant mahogany enamel. **$39.95* List**

**Prices slightly higher west of the Rockies.*

61 SHERMAN ST., MALDEN, MASS.

COMMERCIAL EQUIPMENT

Designed and Built to Exacting Specifications

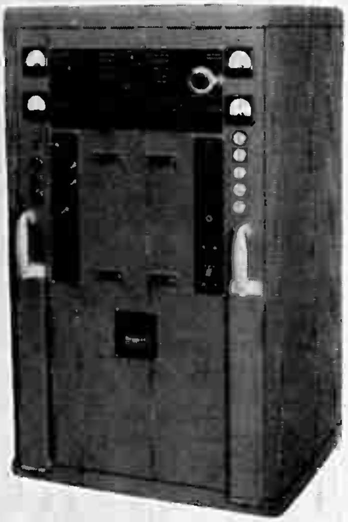
Over the years, the National Company has been called upon by the government and industry to design and build radio and electronic equipment for specialized purposes. National engineers have met this challenge by producing equipment that, in all cases, has *more than met* the most exacting specifications . . . equipment that is operating dependably today all over the world. A few examples are illustrated at the right.

If you use or need similar equipment, why not write National today for further information. Address inquiries to the Commercial Division.

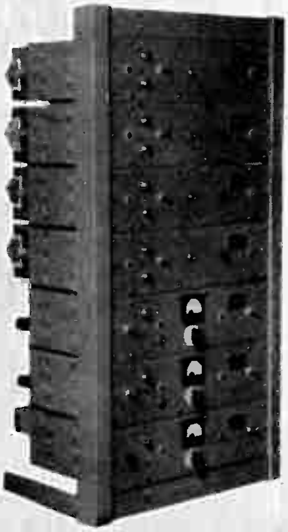
National
©
EST. 1914
NATIONAL COMPANY, Inc.
MALDEN, MASSACHUSETTS

EXPORT INQUIRIES on all National products — television, communication receivers, commercial equipment and components — should be addressed to Export Div., Dept. HB.

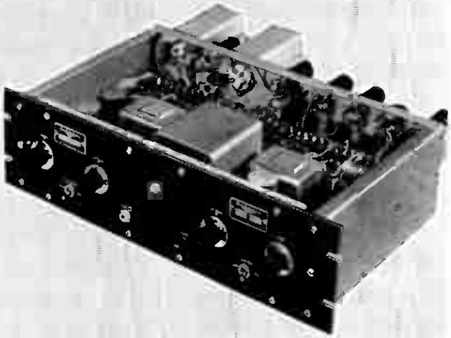
NATIONAL COMPANY, INC.
61 Sherman St., Malden, Mass.



VHF Transmitter



Frequency Shift Equipment



Single-Channel Receiver

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 moulded R-39 plug has two banana plugs on ¾" 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 ½" 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, ¼" diam., 32 thread.

XS-7,	(⅜" Hole)	Net $.36
XS-8,	(½" Hole)	Net $.48
XS-1,	(1" Hole)	Net $.72
XS-2,	(1½" 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 **Net $.36**
A low-loss steatite spreader for 6 inch line spacing. (600 ohms impedance with No. 12 wire.)

AA-5 **Net $.30**
A low-loss steatite aircraft-type strain insulator.

AA-6 **Net $.54**
A general purpose strain insulator of low-loss steatite.

GS-1,	½" x 1⅜"	Net $.24
GS-2,	½" x 2⅞"	Net $.30
GS-3,	¾" x 2⅞"	Net $.60
GS-4,	¾" x 4⅞"	Net $.75

GS-4A, ¾" x 6⅞"
Net $1.05

Cylindrical low-loss steatite standoff insulators with nickel plated caps and bases.

GSJ, (not illustrated)
Net $.10

A special nickel plated jack top threaded to fit the ¾" diameter insulators GS-3, GS-4 & GS-4A.

GS-10, ¾" high
Net, box of ten $.90

GS-10S (not illustrated) but same as GS-10 except includes threaded stud in top end. **Net, box of ten $1.00**

GS-5,	1¼" high	Net $.30
GS-6,	2" high	Net $.42
GS-7,	3" high	Net $.75

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 ¼" deep and GS-10S as noted above.

GS-8, with terminal **Net $.54**
GS-9, with jack **Net $.75**
These low-loss steatite standoff Insulators are also useful as lead-through bushings.

XS-3, (2¾" hole) **Net $3.60**
XS-4, (3¾" hole) **Net $4.35**
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
Net, each $ 4.95

XS-5F, With Fittings
Net, per pair $10.20
These big low-loss bowls have an extremely long leakage path and a 5¼" flange for bolting in place. Insulation steatite. Fittings include nickel plated brass spindles, lugs, nuts and washers.

NATIONAL COMPANY, INC., 61 SHERMAN ST., MALDEN, MASS.

POPULAR National COMPONENTS


HRT


R

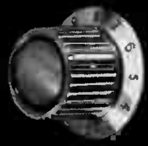

HRP-P


HRS-1


HRP


HRS-2


HRK


HRS-3

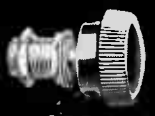

HR


HRT-M

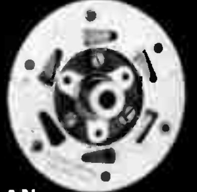

HRB

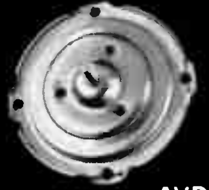

SB


ODL


ODD


RSL


AN


AVD

HRT (gray or black) **Net $.75**

The HRT knob is 2⅛" in dia. and fits ¼" 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 1⅜" dia. chrome plated bevel skirts fit ¼" 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
HRS-5	0-10	through 180°

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 ½" dia. (Fits ¼" 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.

RSL (fits ¼" shaft) **Net $.57**
Rotor shaft lock for TMA, TMC and similar condensers.

AN Vernier Mechanism **Net $**

A vernier mechanism ratio 5-1 an insulated output shaft coup for ¼" shafts. Drive Shaft 3/16" knob.

AVD Vernier Mechanism **Net $**

Similar to AN-Output shaft co ling is non insulated.
For commercial uses many varia available. Write for further ticulars.

R **Net $**

This small dial has a 1⅝" scale calibrated 0-10 in 180° increased reading with clock rotation. Black bakelite knob. ¼" shaft.

HRP-P **Net $**

Black bakelite knob 1¼" long ½" wide. Equipped with poi Especially suitable for use on w and other rotary switches on oratory equipment and the (Fits ¼" shaft).

HRP **Net $**

The type HRP knob has no poi but is otherwise the same as knob above. Recommended for calibrated or hard-tuning cont (Fits ¼" shaft).

HRK **Net $**

Black bakelite knob 2⅜" dial extremely rugged. This is the k used on National type O and L dials.

HRT-M **Net $**

This is a smaller version of the and was designed originally use on the NC-57 Receiver — available in choice of gray or b — is 1-7/16" in diameter.

NATIONAL COMPANY, INC., 61 SHERMAN ST., MALDEN, MASS.

POPULAR National COMPONENTS

ial Net $4.50
Dial Net $3.00
four-inch N and AD Dials have
ne divided and die stamped
es respectively. The N Dial has
ecimal vernier; the AD Dial em-
s a pointer. The planetary drive
a ratio of 5 to 1, and is con-
ed within the body of the dial.
, 4, 5 or blank scale. Fits ¼"
t. **Specify scale.**

ial Net $2.70
vet Vernier" Dial, Type B, has a
pact veriable ratio 6 to 1 min.,
to 1 max. drive that is smooth
trouble free. The case is black
lite. 1 or 5 scale. 4" dia. Fits
shaft. **Specify scale.**

Dial Net $2.10
BM Dial is a smaller version
he B for use where space is limi-
The drive ratio is fixed. Al-
igh small in size, the BM Dial
the same smooth action as the
er units. 1 or 5 scale. 3" dia.
¼" shaft. **Specify scale.**

Dial Net $2.25
original "Velvet Vernier" mech-
m in a metal skirted dial 3" in
ratio 5 to 1. It is available
2, 3, 4, 5 or 6 scale and fits
shaft.

ial Net $1.00
new P dial is the same as the
except direct drive.
e O, 3½" dia., scale 2, with
knob, fits ¼" shafts. Net $1.00
e L, same as O except 5" dia.,
e 2 only. Net $1.95
e K, same as O except less knob,
plete with ODD vernier drive,
e 2 only. Net $1.50
e M, same as K except 5" dia.,
e 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 3½" 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 ¼" below the dial frame. 2¾" H. x 3⅞" 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 6¼" W.

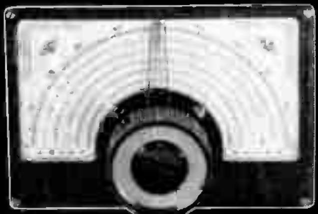
ICN Dial **Net $6.00**

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. 5⅛" H. x 7¼" W.

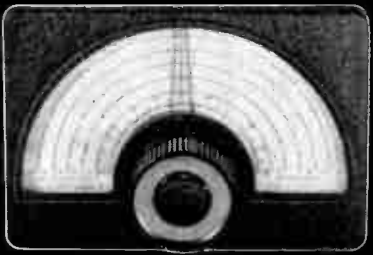
ACN Dial **Net $3.30**

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 7¼" W.

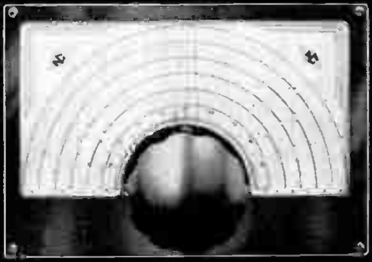
MCN



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
L

K
M

N
AD

B

BM

P

POPULAR National COMPONENTS

XLA

TSA-1

TSA-2

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.

TURRET SOCKET ASSEMBLIES

TSA-1, TSA-2 Designed for our 7-pin and 9-pin miniature tube sockets. Permits compact sub-assembly wiring at base of socket. Cadmium-plated brass center support has a standard length of two inches. Silver-plated brass terminal studs. Available either with holes through which leads can be drawn, or with solid studs. Center supports of varying lengths and other types of terminals can be supplied to manufacturers in quantity.

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

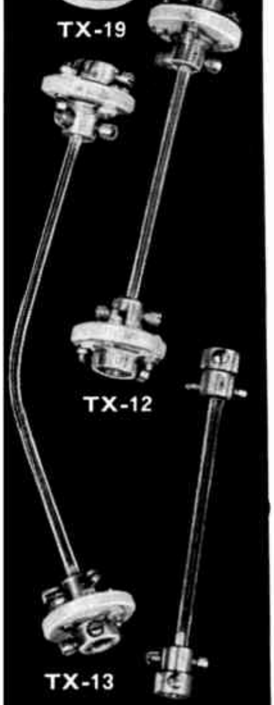
HX-29

XM-10

HX-100S

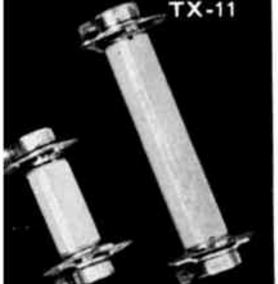
POPULAR *National* COMPONENTS


TX-19


TX-12

TX-13

TX-11


TX-1

TX-2


TX-8


TX-23


TX-9


TX-10

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 collars 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. 1⅜" 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 offers 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 on 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 or 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 other 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
		Max.	Min.	
HC-26	36¢	.051	.045	3C24, HK24, 24G, 25T
HC-27	36¢	.062	.058	UH50, 304B, 829B, 832A, 834
HC-28	36¢	.072	.062	35T, 35TG, 75TH, 8001
HC-29	50¢	.126	.120	152TH
HC-30	50¢	.365	.350	4-125A, 250R, 250TH, 25TL, 802, 804, 807, 814, 815, 828
HC-31	60¢	.128	.116	304TH, 304TL
HC-32	60¢	.573	.558	100R, 450TH, 803, 805, 806, 808, 809, 810, 811, 812, 813, 833A, 866, 1500T, 2000T, 8000, 80003, 80005, HF100, ZB60, HF60, 111H, 211H, 203H, HF175, 5311, 5332
HC-33	80¢	.807	.793	WL460, WL463, WL468, HF200, HF201, HF300


SPP-9


SPP-3


12


24


8


ACD-1

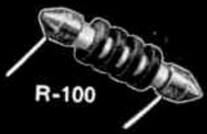
ACD-2

RIGHT ANGLE DRIVES


HEAT RADIATING CAP

NATIONAL COMPANY, INC., 61 SHERMAN ST., MALDEN, MASS.

POPULAR National COMPONENTS



R-100

R-100U

R-100S

R-100ST

R-33

R-50

R-50

R-33G

R-60

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.

R-300

R-300U

R-300S

R-300ST

R-152

R-154

R-154U

R-175

NATIONAL COMPANY, INC., 61 SHERMAN ST., MALDEN, MASS.

POPULAR *National* COMPONENTS

I. F. TRANSFORMERS

IFC
IFCO

IFL
IFM
IFN
IFO

OSR

XR-50

AR-5

AR-2

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**

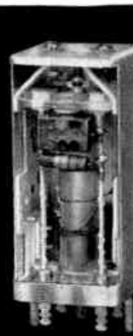
15 Mc. IF transformers suitable for ultra high frequency superheterodynes. They are made in two models with and without variable coupling. Approximate stage gain of 10 is obtained with IFJ or IFK Transformer and 6AB7 tube.

SA:4842 **Net $4.50**

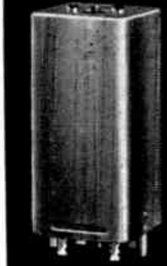
A 456 kc. discriminator transformer for narrow band frequency modulation. This unit is the nucleus of the NFM adapter described by Harrington and Bartell in November 1947 QST. Two slug-tuned secondaries are employed and discrimination is accomplished by resonating one at approximately 10 kc. above, the 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 H.F. Coil **Net $1.70**
AR-5 H.F. 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.

XR-51 same but with brass slug

OSR **Net $1.80**

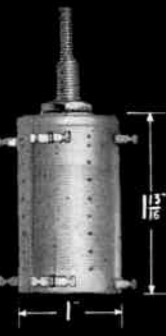
A shielded oscillator coil which tunes to 100 kc. with .00041 mfd. Two separate inductances, closely coupled. Excellent for interruption-frequency oscillator in superregenerative receivers.

CERAMIC SLUG-TUNED COIL FORMS

XR-70 (grooved for #19 wire, with iron slug) **Net $1.32**
XR-71 (same, brass slug) **Net $1.32**
XR-72 (not grooved, winding length 1", with iron slug) **Net $1.32**
XR-73 (same, brass slug) **Net $1.32**
XR-60 (grooved for #26 wire, with iron slug) **Net $1.32**
XR-61 (same, brass slug) **Net $1.32**
XR-62 (not grooved, winding length 1¼", with iron slug) **Net $1.32**
XR-63 (same, brass slug) **Net $1.32**

High-grade ceramic coil forms conforming to JAN specifications. May be wound as desired to provide a permeability-tuned coil. Extra lugs provided.

XR-70 XR-71
XR-72 XR-73



XR-60 XR-61
XR-62 XR-63

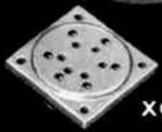
NATIONAL COMPANY, INC., 61 SHERMAN ST., MALDEN, MASS.

POPULAR National COMPONENTS

XR-1
XR-2

XR-3

XR-4
XR-5
XR-6



XC-6C

SC-1

SC-4

CFA

PH-1

PLUG-IN BASE AND SHIELD

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 ⅛" and 3/32" and ⅛" 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 he 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**
1⅞" 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)

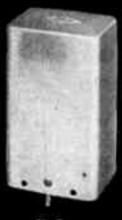
CP-1, dark gray **Net $.40**

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**

RZ

RS



RO

T-78

JS-1

XOS-1

XOS-

XOS-3

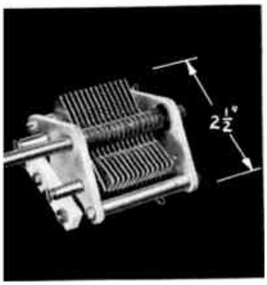
FIXED-TUNED EXCITER TANK

NATIONAL COMPANY, INC., 61 SHERMAN ST., MALDEN, MASS.

POPULAR National COMPONENTS

TYPE TMS TRANSMITTING CONDENSERS

s is a condenser designed for transmitter use in low power stages. It is compact, rigid, and dependable. Provision has en made for mounting either on the panel, on the chassis, or on two stand-off insulators. Insulation is steatite. Voltage ra- s listed are conservative.

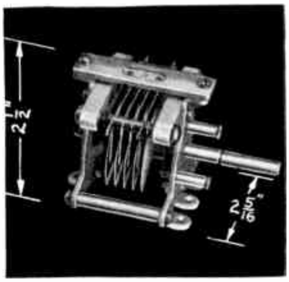


Capacity	Minimum Capacity	Length	Air Gap	Peak Voltage	No. of Plates	Catalog Symbol	Net
SINGLE STATOR MODELS							
100 Mmf.	9.5	3"	.026"	1000v.	9	TMS-100	$2.60
150	11	3"	.026"	1000v.	14	TMS-150	2.80
250	13.5	3"	.026"	1000v.	22	TMS-250	3.30
300	15	3"	.026"	1000v.	27	TMS-300	3.80
35	8	3"	.065"	2000v.	7	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

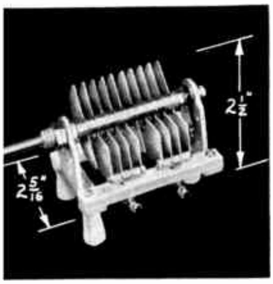
is is a new condenser for exciters and low power transmitters. Special provision has been made for mounting AR-16 coils a swivel plug-in mount on either the top or rear of the condenser. For stand-off or panel mounting-steatite insulation.

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 3/8"	.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
100–100	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 olished edges. It mounts on the panel or on removable stand-off insulators. Steatite insulators have long leakage path. tand-offs included in listed price.

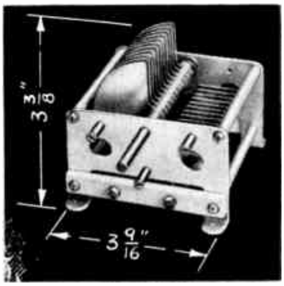


Capacity	Minimum Capacity	Length	Air Gap	Peak Voltage	No. of Plates	Catalog Symbol	Net
SINGLE STATOR MODELS							
50 Mmf.	9	3 3/4"	.085"	3500v.	15	TMH-50	$3.95
75	11	3 3/4"	.085"	3500v.	19	TMH-75	4.15
100	12.5	5 1/8"	.085"	3500v.	25	TMH-100	4.35
150	18	6 1/2"	.085"	3500v.	37	TMH-150	4.95
35	11	5 1/8"	.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

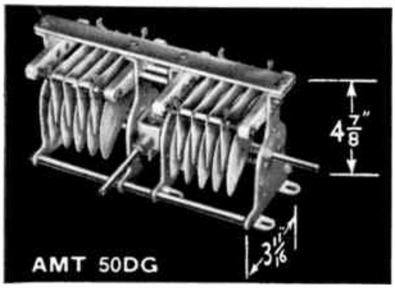
condenser designed for use in the power stages of transmitters where peak voltages do not exceed 3000 volts. The frame is xtremely rigid and arranged for mounting on panel, chassis or stand-off insulators. The plates are aluminum with buffed dges. Insulation is steatite. The stator in the split stator 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 3/8"	.077"	3000v.	39	TMC-300	6.10
DOUBLE STATOR MODELS							
50–50 Mmf.	9–9	4 5/8"	.077"	3000v.	7–7	TMC-50D	$4.35
100–100	11–11	6 3/4"	.077"	3000v.	13–13	TMC-100D	5.95
200–200	18.5–18.5	9 1/4"	.077"	3000v.	25–25	TMC-200D	7.25



NATIONAL COMPANY, INC., 61 SHERMAN ST., MALDEN, MASS.

POPULAR National COMPONENTS

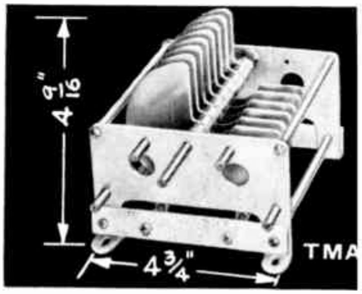


TYPE AMT

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 ¼" dia. shaft extension.



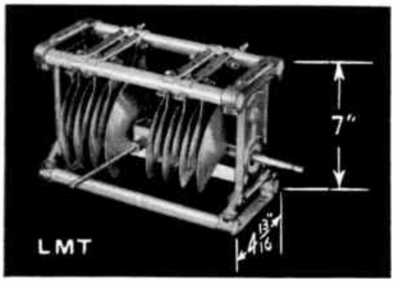
TYPE TMA

This is a larger model of the popular TMC. The frame is extremely rigid and arranged for mounting on panel, chassis or sta 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	4¾"	.177"	6000 v.	9	AMT-50	$ 5.20
100	20	6¾"	.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	6⅞"	.171"	6000 v.	15	TMA-100A	5.85
150	22.5	6⅞"	.171"	6000 v.	21	TMA-150A	6.45
230	33	9 5/16"	.171"	6000 v.	33	TMA-230A	7.95
100	30	9¼"	.265"	9000 v.	23	TMA-100B	8.50
150	40.5	12½"	.265"	9000 v.	33	TMA-150B	9.95
50	21	7⅛"	.359"	12,000 v.	13	TMA-50C	5.55
100	37.5	12⅞"	.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	13⅝"	.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	13⅝"	.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	13⅝"	.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	9⅜"	.177"	6000 v.	18	AMT-50D	7.00
100–100	20–20	13⅜"	.177"	6000 v.	34	AMT-100D	9.00
50–50	13–13	9⅜"	.177"	6000 v.	18	AMT-50DG	10.75
100–100	20–20	13⅜"	.177"	6000 v.	34	AMT-100DG	12.75
200–200	15–15	6⅞"	.077"	3000 v.	16–16	TMA-200D	9.40
180–180	10–10	12¾"	.140"	4000 v.	24–24	TMA-180D	12.90
50–50	12.5–12.5	6⅞"	.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	12½"	.249"	9000 v.	15–15	TMA-60DB	8.95
40–40	18–18	12⅞"	.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	13⅝"	.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

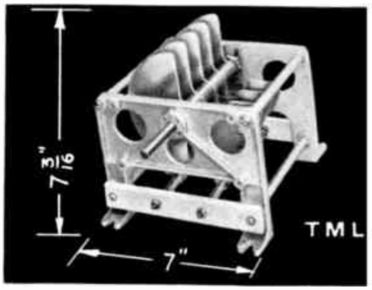
TYPE LMT

A heavy duty transmitting condenser that completely eliminates troublesome closed loops, vastly simplifying the probl of unwanted harmonics. The rotor shaft is completely insulated from the end plates. Long leakage path (higher safety facto Plates and parts are extra heavy with highly polished rounded edges to prevent flash-over. Adjustable stator plate mounti and end bearings. Available in single-stator, double-stator, or double-stator right angle center drive models. Same capacit 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.



NATIONAL COMPANY, INC., 61 SHERMAN ST., MALDEN, MASS.

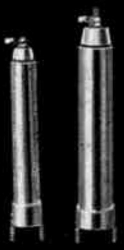
POPULAR National COMPONENTS

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 screwdriver 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**

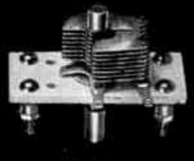
STN **Net $2.07**
The Type STN has a maximum capacity of 18 mmf. (3000 V), making it suitable for such tubes as the 809. It is supplied with two standoff insulators.

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.



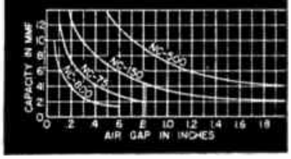
UM

UMA-25

UMB-25

NC-800A

NC-75
NC-150
NC-500



POPULAR 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 ¼" 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**

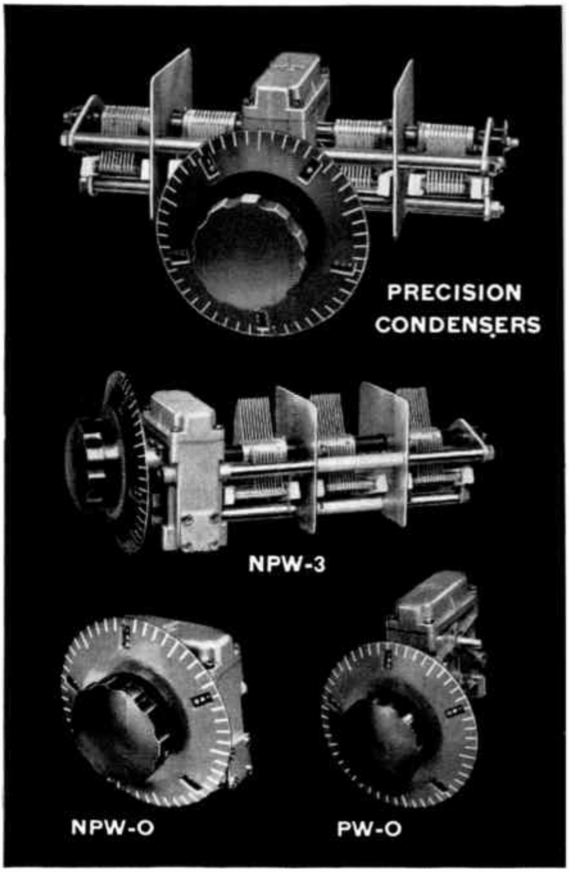
Similar to PW models, except that rotor shaft is perpendicular to panel.

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 tl complete range and as there is no gear reduction unit furnished, the driven shaft will revolve ten times, also. The PW- 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; tl coils are never changed. The unit is built around a circuit which tunes to two harmonically unrelated frequencies at the san time. Thus, it becomes possible to cover a wide frequency range and yet maintain a reasonably constant L/C ratio. 3" wic x 8¼" high (including the GS-10 standoffs) x 9" long overall including the ¼" dia. shaft and output terminals.

MB-40L

Features of the MB-150:

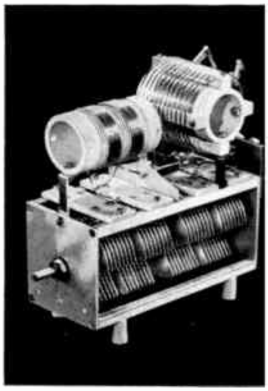
(1) For use as the all-band plate tank in push-pull or single-ended stages running up to 150-watts input (1500 volts peak). It is ideal for a pair of 807s or 809s or a single 829B.

(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

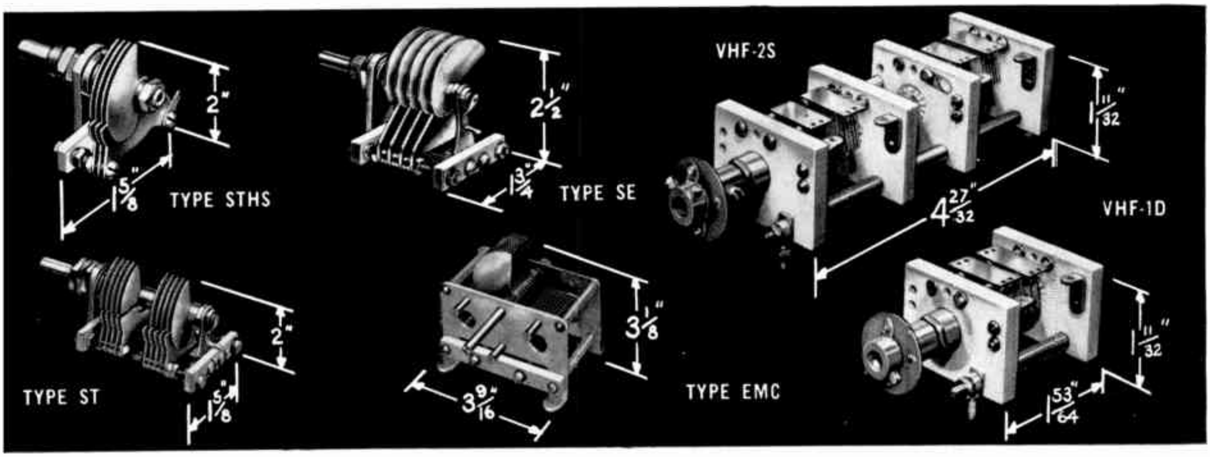


MB 40L LOW-POWER MULTI-BAND TANK

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 $12.90**

NATIONAL COMPANY, INC., 61 SHERMAN ST., MALDEN, MASS.

POPULAR National COMPONENTS



TYPE ST (180° Rotation)
STRAIGHT-LINE WAVELENGTH

: ST Type condenser has Straight-Line Wavelength plates. All double-
aring models have the front bearing insulated to prevent noise. On special
ler a shaft extension at each end is available, for ganging. On double-
aring single shaft models, the rotor contact is through a constant impedance
tail Steatite insulation.

DTE — Type SS Condensers, having straight-line capacity plates but
ierwise similar to the Type ST, are available. Capacities and Prices same
Type ST.

apacity	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
			SPLIT STATOR DOUBLE BEARING MODELS			
50–50	5–5	11–11	.026"	2¾"	STD- 50	$3.60
00–100	5.5–5.5	14–14	.018"	2¾"	STHD-100	3.90
			DOUBLE BEARING MODELS			
35 Mmf.	6 Mmf.	8	.026"	2¼"	ST- 35	$1.85
50	7	11	.026"	2¼"	ST- 50	1.90
75	8	15	.026"	2¼"	ST- 75	2.00
00	9	20	.026"	2¼"	ST-100	2.10
40	10	27	.026"	2¾"	ST-140	2.30
50	10.5	29	.026"	2¾"	ST-150	2.30
00	12.0	27	.018"	2¼"	STH-200	2.50
50	13.5	32	.018"	2¾"	STH-250	2.70
00	15.0	39	.018"	2¾"	STH-300	2.90
35	17.0	43	.018"	2¾"	STH-335	3.10

TYPE SE (270° Rotation)
STRAIGHT-LINE FREQUENCY

/PE SE — All models have two rotor bearings, the front bearing being
sulated to prevent noise. A shaft extension at each end, for ganging, is
ailable on special order. On models with single shaft extension, the rotor
ntact is through a constant impedance pigtail. The SEU models (illustrated)
e suitable for high voltages as their plates are thick polished aluminum with
unded edges. Other SE condensers do not have polished edges on the
ates. Steatite insulation.

15 Mmf.	7 Mmf.	6	.055"	2¼"	SEU- 15	$2.80
20	7.5	7	.055"	2¼"	SEU- 20	2.95
25	8	9	.055"	2¼"	SEU- 25	3.10
50	9	11	.026"	2¼"	SE- 50	2.30
75	10	15	.026"	2¼"	SE- 75	2.40
100	11.5	20	.026"	2¼"	SE-100	2.60
150	13	29	.026"	2¾"	SE-150	2.75
200	12	27	.018"	2¼"	SEH-200	2.80
250	14	32	.018"	2¾"	SEH-250	3.00
300	16	39	.018"	2¾"	SEH-300	3.25
335	17	43	.018"	2¾"	SEH-335	3.50

TYPE EMC (180° Rotation)
STRAIGHT-LINE WAVELENGTH

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 frame 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	4⅜"	EMC- 500	6.75
1000	22	56	6¾"	EMC-1000	10.35

VHF CONDENSERS

● Shaft extension at rear for ganging purposes. Dual condensers ideal for mixer-oscillator unit. ● Ball bearings front and back for smooth rotation and freedom from back-lash. ● Brackets for mounting 7-pin miniature tube sockets, i.e., National XOA for very short leads from tube to condenser essential for VHF efficiency, and rigid compact unit-assembly that produces better stability. ● Wide low-inductance stator strap connections raise frequency limit of condensers. Coil or strap tank can be connected directly to stator straps allowing maximum inductance in tank and a minimum of inductance between tank and stator. ● Stators, rotors and stator strap connections silver-plated for best efficiency. ● Rigid square construction, heavy isolantite end plates. ● Spade bolts allow solid connections to chassis for extreme rigidity. ● Flexible insulating coupling available to connect condenser shaft to ¼" dial shaft. ● Flexible insulating coupling available to connect two or more condensers together as ganged units. ● High capacity single spaced units for general coverage. ● Low capacity double spaced units for bandspread, suitable for ham use, particularly in the VHF and UHF ham bands. ● Stators solder construction can be removed and replaced by strap tanks for special VHF and UHF application.

DOUBLE SPACED MODELS

Two section VHF-2D, price $6.50.

Maximum capacity per section stator to stator	6.75 mmf.
Minimum capacity per section stator to stator	3.0 mmf.
Net change	3.75 mmf.

Single section VHF-1D, price $3.25.

Maximum capacity stator to stator	6.75 mmf.
Minimum capacity stator to stator	3.0 mmf.
Net change	3.75 mmf.

SINGLE SPACED MODELS

Two section VHF-2S, price $6.50.

Maximum capacity per section stator to stator	22.5 mmf.
Minimum capacity per section stator to stator	3.0 mmf.
Net change	19.5 mmf.

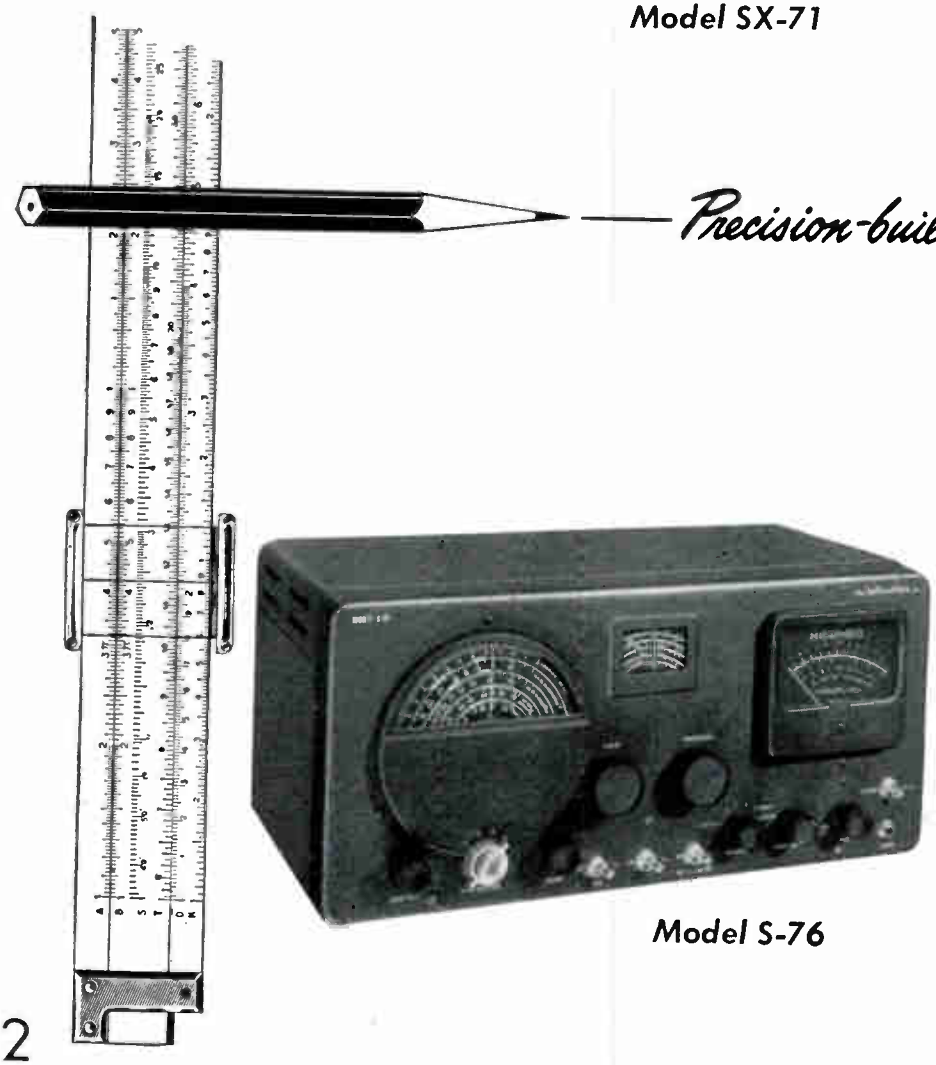
Single section VHF-1S, price $3.25.

Maximum capacity stator to stator	22.5 mmf.
Minimum capacity stator to stator	3.0 mmf.
Net change	19.5 mmf.

NATIONAL COMPANY, INC., 61 SHERMAN ST., MALDEN, MASS.

hallicrafters

Model SX-71



Model S-76

For Definitely Superior Ham Performance

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. Extra wide dials for main and bandspread tuning. Surpasses in ham performance many receivers priced considerably higher.

PERFORMANCE: Continuous AM reception from 538 kc to 35 Mc, and 46 to 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. Temperature compensated, voltage regulated. One r-f, two conversion, and 3 i-f stages yield high gain for sensitivity of .7 microvolts with 50 milliwatts output. Audio peaked for communications frequencies, with 3 watt output.

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 controls; bandspread dial calibrated for 80, 40, 20, 10, and 6 Meter Bands. BFO Pitch 3-position Selectivity, Crystal Phasing, Tone, a-f Gain, and r-f Gain controls. ANL, BFO, and Receive/Send switches. "S" meter adjustment on rear.

PHYSICAL DATA: Satin black steel cabinet with chrome trim. Piano hinge top. Size 18½ in. wide by 8⅞ in. high by 12 in. deep. Ship. wt. 51 lbs.

EXTERNAL CONNECTIONS: Use doublet or single wire antenna. 500 and 3.2 ohm outputs for separate speaker. Phone jack. Socket for external power supply. Connections for remote control. For 105—125 volts 50/60 cycle AC.

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 Reg., and 5Y3GT Rect.

UNIVERSAL MODEL SX71U: Same as above only for 115/250 volts, 25/60 cycle AC.

R-46 SPEAKER. New 10" PM in satin black steel cabinet to match SX-71 and S-76 (also suitable for SX-62). 500-ohm transformer. 80 to 5000 cycle range. 15" wide, 10⅞" high, 10⅞ deep.

communications equipment

New! A Double Superhet With 50 kc I-F

A new double conversion receiver just introduced as the lower-priced running mate to the already famous SX-71. The only double superhet with 50 kc second i-f and the only set now known with a giant sized 4-inch "S" Meter. Another new Hallicrafters engineering triumph . . . a special value leader in the moderate price range.

PERFORMANCE: Continuous coverage 538-1580 kc and 1.72-32 Mc. Double conversion almost completely eliminates images. 50 kc second i-f gives excellent "skirt" selectivity with "nose" selectivity variable from 5.0 kc down to 500 cycles. Temperature compensated, voltage regulated. One r-f, two conversion, and two i-f stages. 2½ watts output, with audio peaked for communications frequencies.

CONTROLS: Band Selector 538-1580 kc, 1.72-4.9 Mc, 4.6-13 Mc, 12-32 Mc; Separate Main and Bandspread tuning; bandspread calibrated for 80, 40, 20, 11, 10 meters; five-position Selectivity with phono switch built-in; BFO Pitch; full-range Tone; AVC, BFO, ANL, Rec./Standby switches. "S" Meter adjustment on rear.

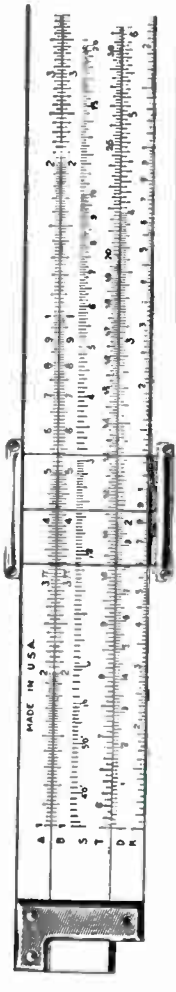
PHYSICAL DATA: Satin black steel cabinet with chrome trim. Piano hinge top. Size 18½" wide, 8⅞" high, 9" deep. Ship. wt. approx. 46 lbs.

EXTERNAL CONNECTIONS: Use doublet or single wire antenna. 500 or 3.2 ohm outputs. Phone jack. Phono input jack. Connections external power and for remote control. Mounting holes provided for coax connector. For 105-125 volts 50/60 cycle AC.

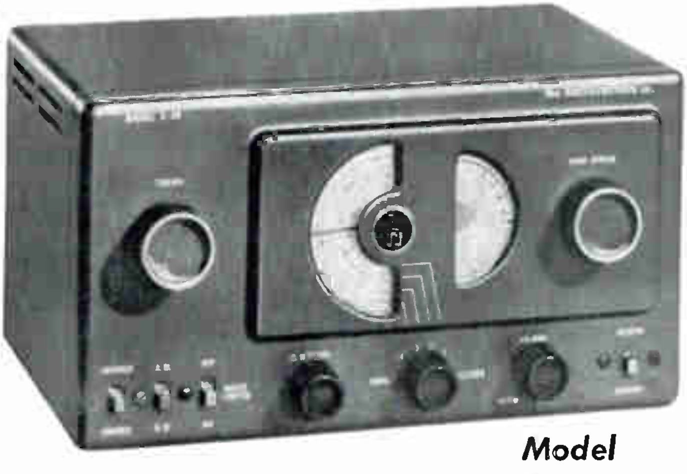
9 TUBES PLUS REGULATOR AND RECTIFIER: 6CB6 r-f Amp., 6AU6 1st Conv., 6C4 Osc., 6BA6 1st i-f 6BE6 2nd Conv., 6BA6 2nd i-f, 6AL5 Det., ANL, 6SC7 BFO, 6K6GT Output, VR-150 Reg., 5Y3GT Rect.

hallicrafters

Model
S-40B & S-77



Model
S-53A



Model
S-38B

New Versions of an Old Favorite

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 chrome trim. Top opens on piano hinge. Size 18½ in. wide by 8⅞ 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. S-40B uses 105-125 V. 50/60 cycle AC only. S-77 uses 105-125 V. DC or 50/60 cycle AC.

7 TUBES PLUS RECTIFIER: (in S-40B) 6SG7 RF Amp., 6SA7 Conv., two 6SK7 IF Amps., 6H6 ANL and AVC, 6SL7 BFO and Det., 6F6G Output, 5Y3GT Rectifier. Comparable AC/DC type tubes used in S-77.

UNIVERSAL MODEL S-40BU: Same as above only for 115/250 volts, 25/60 cycle AC.

Superb Performance in Compact Size

Unquestionably the finest small communications receiver built. Several steps better than the S-38B but not as good as the S-40B. Complete in itself, with built-in PM speaker.

PERFORMANCE: Coverage 540—1600 kc, 2.6—31 Mc plus 48—54.5 Mc Two stages IF amplification.

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 12⅞ in. wide by 7 in. high by 7¾ 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.

UNIVERSAL MODEL S-53AU: Same as above only for 115/250 volts, 25/60 cycle AC

The Radio That Amazes the Experts

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 12⅞ in. wide by 7 in. high by 7¼ 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.

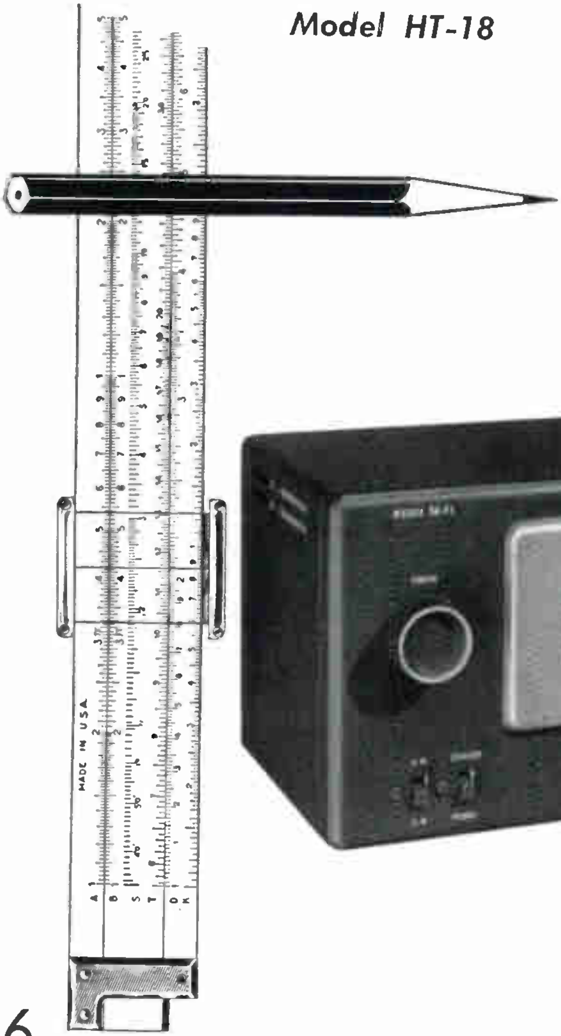
4 TUBES PLUS RECTIFIER: 12SA7 Conv., 12SK7 IF Amp. and BFO, 12SQ7 Det. and AVC, 35L6GT Output, 35Z5GT Rectifier.

220-VOLT LINE CORD: Available separately. Works for AC or DC.

"The Radio Man's Radio"

hallicrafters

Model HT-18



Precision-built

Model SR-75

Popular Calibrated VFO with 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. Consists of an oscillator (crystal controlled or VFO), a frequency modulator with speech amplifier, and a buffer-output tube.

CONTROLS: Operatian Switch has three crystal positions plus VFO and NBFM; Band Selector switch — 80, 40, 20, 15, 10 Meters; 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. Tuning control operates osc. gang and calibrated dial.

PHYSICAL DATA AND CONNECTIONS: Satin black steel cabinet with brushed chrome trim; size 12⅞" wide, 7" high, by 7¾" deep. Shipping weight 24 lbs. Connections for microphone, keying (osc. keying), remote control, and 72-ohm output. Line cord for 115 V. 50/60 cycle AC.

TUBES: Three 6BA6—Osc., Freq. Modulator, Speech Amp., 6L6 Buffer, VR-105 Voltage Reg., 5Y3GT Rectifier.

communications equipment

New Transceiver with Amazing Performance

A completely new type of unit—a small transceiver for the novice class and/or beginning amateur; can also be used later as exciter unit. Receives on 540 kc through 32 Mc, transmits on 10, 11, 20, 40, or 80 meter bands. 10 watts input to final amp.

Receiving section is substantially same as our S-38B. Bandspread tuning, Speaker/phones switch, BFO switch, Rec. Standby switch; four tubes plus rectifier.

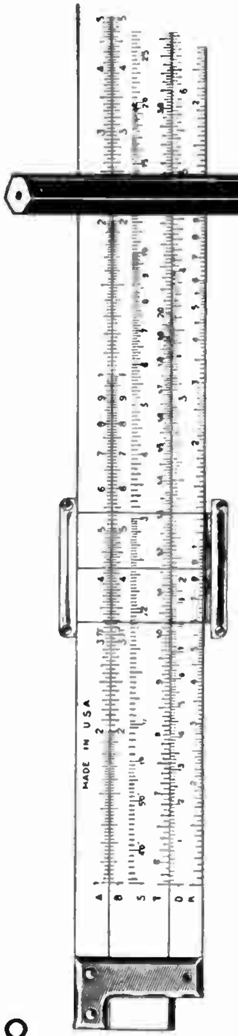
Transmitting section uses electron coupled Xtal oscillator plus output tube of receiver. Oscillator keying, through relay, so completely isolated. Voltage doubler rectifier to increase plate voltage.

CONTROLS: Main Tuning in Mc; separate electrical bandspread; AM/CW, Speaker/Phones, and Receive/Standby switches, Volume control with power switch. Xmtr controls on rear; tuning, coupling adj., doubler coil switch (10 meters) and adjustment, power switch with interlock.

PHYSICAL DATA AND CONNECTIONS: Satin black steel cabinet with brushed chrome trim; size 12⅞" wide, 7" high, by 7¾" deep. Shipping weight 16 lbs. Connections for keying, headphone tip jacks, tuning meter or bulb, and output. Line cord for 115 volts 50/60 cycle AC or DC. Shipped with coils, less crystals.

hallicrafters

Model SX-62



Precision-built

Model S-72 & S-72L

Designed for Top Broadcast Reception

The world's finest receiver for the All-Wave listener. 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 a fine 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 10¼ in. high by 16 in. deep. Ship. Wt. 66 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.

UNIVERSAL MODEL SX-62U: Same as above only for 115/250 volts, 25/60 cycle AC

communications equipment

Regular and Long-Wave 3-Way Portable

You'll always be in touch with the outside world wherever you go with this new Hallicrafters extra-sensitive portable. Designed both for the person who wants better than average operation even in weak signal areas and for the Radio Amateur.

PERFORMANCE: Regular Model S-72 covers standard broadcast and three short-wave bands 540 kc to 30 Mc continuously. Long-Wave Model S-72L covers airways ranges and towers and marine beacons 175-420 kc, plus Broadcast and 2 short-wave bands 540 kc to 12.5 Mc. One stage tuned r-f amplification; separate bandspread tuning gang. Two built in antennas—loop for broadcast and 61-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. 18 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 is answer to airplane or boat owner's dream. Receives marine beacons, airways ranges and towers, as well as airways and marine shortwave channels and regular broadcast band.

hallicrafters

Model SX-73

Unsurpassed Performance Under All Conditions

Coming soon from Hallicrafters world-famous short wave laboratories will be a superb new communications receiver—the SX-73, proud successor to so many famous top-quality Hallicrafters receivers. Absolutely without equal in its combination of ruggedness, sensitivity, stability, selectivity, resetability, and image and i-f rejection. Based on an original design developed by Hallicrafters for the armed forces for universal use all over the world, this new receiver will surpass all others in versatility, dependability, performance and value.

PERFORMANCE: Continuous frequency coverage 540 kc to 54.0 Mc. Two r-f, two i-f stages. Dual conversion above 7 Mc; second beat oscillator is crystal controlled. Choice of six pretuned crystal controlled channels in range 1.5 to 30 Mc. Single tuning knob turns main and bandspread dials (6 to 1 ratio between the two); 50 to 1 tuning ratio.. Resetability accurate to within 30 cycles per megacycle. Selectivity variable 14.5 kc to 300 cycles at 6 db down. Sensitivity less than 2 microvolts for .5 watts output. Signal to noise ratio 10 db for 2 mv input. Image rejection 80 to 120 db. I-f rejection not less than 60 db. AVC circuit will hold up to one volt without overload. Series type noise limiter. Carrier level meter. Audio response plus or minus 1½ db from 300 to 3500 cycles.

CONTROLS: Tuning knob with dial lock; Band Selector 540-1350 kc, 1.35-3.45 Mc, 3.45-7.00 Mc, 7.00-14.4 Mc, 14.4-29.7 Mc, 29.7-54.0 Mc; r-f Gain and AC on/off, BFO Pitch, Xtal Phasing, 6-pos. Xtal Selectivity, 6-pos. Xtal fixed-frequency channel selector, a-f Gain, Xtal tuning Vernier; Rec./Standby, BFO, AVC, and ANL switches; BFO injection control and carrier meter adj. on rear.

PHYSICAL DATA: Two-tone gray steel cabinet with satin chrome trim. Piano hinge top. Size 20 in. wide, 11 in. high, 18½ in. deep.

EXTERNAL CONNECTIONS: Antenna Input 50 to 200 ohms throughout tuning range. Output 600 and 50 ohms. For 50/60 cycle current at 75, 105, 117, 130, 190, 210, 234, or 260 volts.

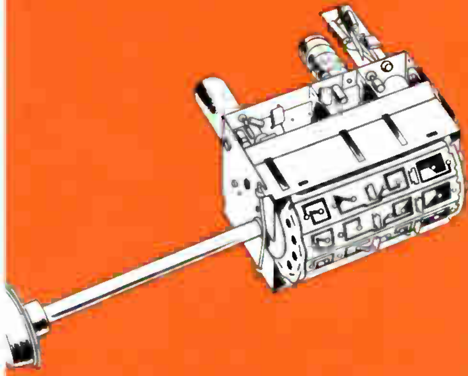
17 tubes plus voltage regulator, ballast tube and rectifier.

Precision-built **communications equipment**

hallicrafters

Precision-Built Television

with the



dynamic tuner

A tuning system more accurate and more powerful than any in Television History

The Dynamic Tuner — a rotary-type tuner — uses flat tuning coils that are precision-printed by a special photo-etch process. Because wire stretches as it is wound, and because coil forms vary, NO OTHER TUNING SYSTEM can even approach the absolute accuracy of precision photo-printed coils.

The heart of the Dynamic Tuner lies in the 12 channel strips. Each strip has been prepared by photographically printing the desired pattern on copper and then etching away the unwanted metal. The result of this process is complete uniformity in production. Every chassis coming off the line is "hot" in sensitivity; variations in tuning alignment are practically eliminated.

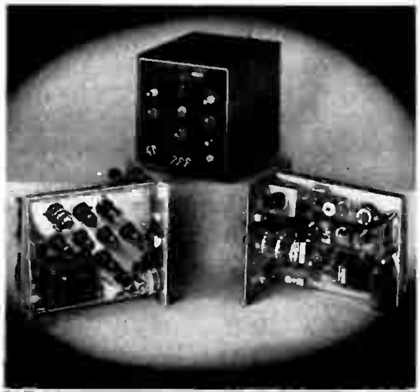
Only Hallicrafters has the Dynamic Tuner, to bring you the clearest picture in television. "City Clear", even in weak signal areas. Independent research laboratories report that Hallicrafters chassis delivered 2 to 4 times more sensitivity than the best of four other leading sets tested. See your Classified Telephone Directory for your nearest Hallicrafters TV dealer.

hallicrafters

Magnificent **IN PERFORMANCE... *Modest* IN PRICE**

THE HALLICRAFTERS CO., CHICAGO 24

JAMES MILLEN
MALDEN · MASSACHUSETTS




90651


90902

90903

90905


90921

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 . . . . $

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 . . . . $

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 . . . . $

Additional Inductors for Lower Frequencies

No. 46702—925 to 2000 KC . . . . $
No. 46703—500 to 1050 KC . . . .
No. 46704—325 to 600 KC . . . .
No. 46705—220 to 350 KC . . . .

LABORATORY SYNCHROSCOPES

The 5" laboratory synchroscopes are available with and without detector-video strips.

Model P-4-2, with tubes . . . . $
Model P-4E-2, with tubes . . . .

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 . . . . $

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 . . . . $
No. 90903, less tubes . . . .
No. 90905, less tubes . . . .

'SCOPE AMPLIFIER—SWEEP UNIT

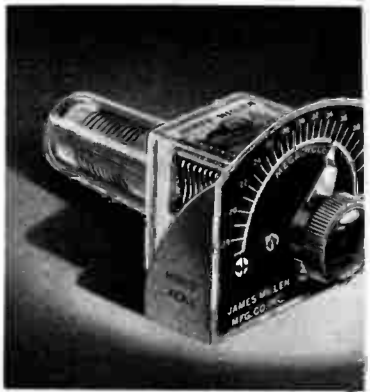
Vertical and horizontal amplifiers along with hard-tube, saw tooth sweep generator. Complete with power supply mounted on a standard 5¼" rack panel.

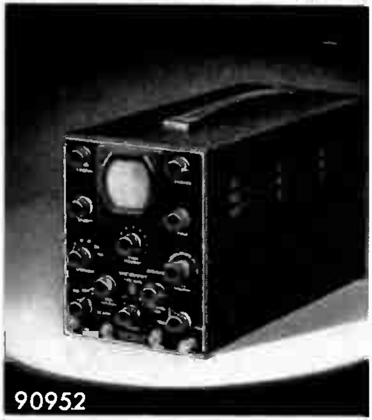
No. 90921, with tubes . . . . $

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 . . . . $




90952


90201

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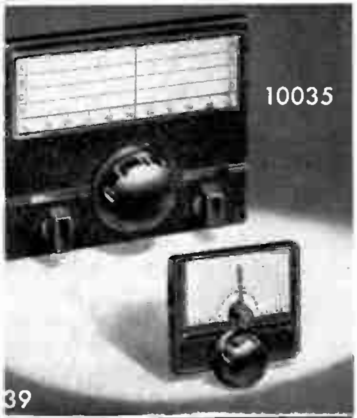
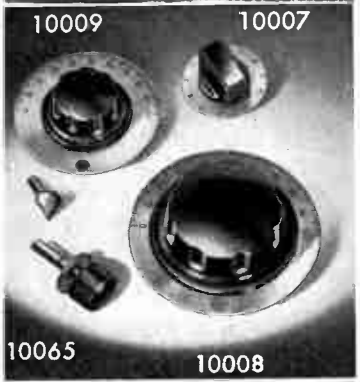


INSTRUMENT DIALS

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 . . . . $
No. 10035 . . . .
No. 10030 . . . .



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 . . . . $
No. 10008 . . . .
No. 10009 . . . .
No. 10021 . . . .
No. 10065 . . . .

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 . . . . $

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 . . . . . . $

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 . . . . $

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 . . . . $

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 . . . . $

JAMES MILLEN
MALDEN · MASSACHUSETTS


92101

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................. $


90671

STANDING WAVE RATIO BRIDGE

The Millen S.W.R. bridge provides easy and inexpensive measurement of standing wave ratio on antennas using co-ax cable. As assembled the bridge is set up for 52 ohm line. A calibrated 75 ohm resistor is mounted inside the case for substitution in the circuit when 75 ohm line is used.

No. 90671............................. $


90700

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............ $


90711

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............... $


90800

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.

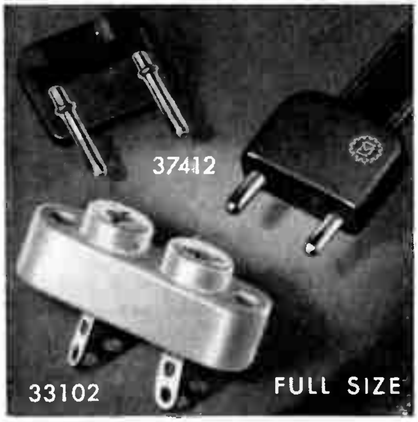
Model 90800, less tubes.............. $


74400

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.

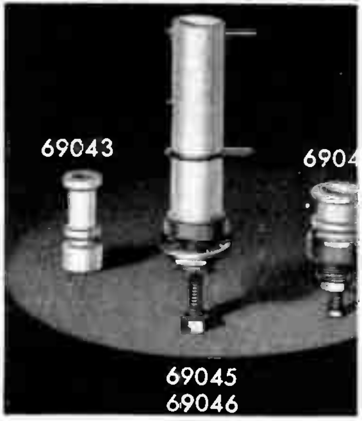
No. 74400.......................... $



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 ½", diameter .095".

No. 37412.......................... $



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; ¼ x ⅜ for 69043–7–8; ½ x 1 1/16 for 69045–6; 3/16 x 3/16 for 69044.

No. 69041—(Copper Slug)............. $
No. 69042—(Iron Core)...............
No. 69043—(Iron Core)...............
No. 69044—(Copper Slug).............
No. 69045—(Copper Slug).............
No. 69046—(Iron Core)...............
No. 69047—(Copper Slug).............
No. 69048—(Iron Core)...............

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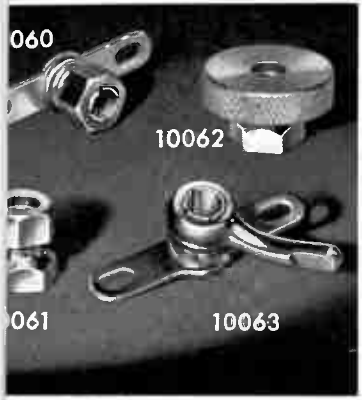


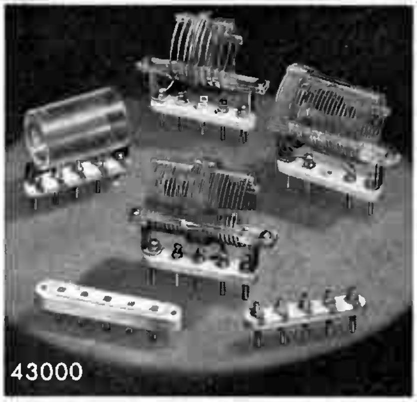
SHAFT LOCKS

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.

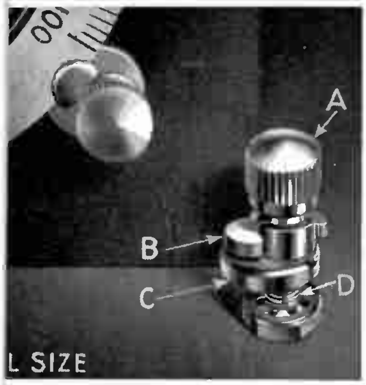
No. 10060 .... $
No. 10061 ....
No. 10062 ....
No. 10063 ....

TRANSMITTING TANK COILS

A full line—all popular wattages for all bands. Send for special catalog.



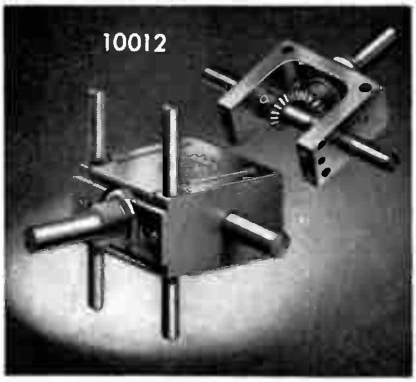
DIAL LOCK



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 .... $

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 .... $

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 .... $

FLEXIBLE COUPLINGS

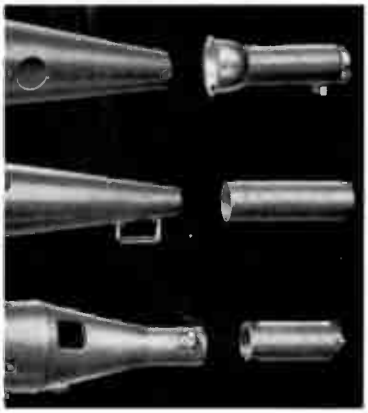
The No. 39000 series of Millen "Designed for Application" flexible coupling units include, in addition to improved versions of the conventional types, also such exclusive original designs as the No. 39001 insulated universal joint and the No. 39006 "slide-action" coupling (in both steatite and bakelite 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 .... $
No. 39002 ....
No. 39003 ....
No. 39005 ....
No. 39006 ....

CATHODE RAY TUBE SHIELDS



For many years we have specialized in the design and manufacture of magnetic metal shields of nicoloi 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—Nicoloi for 5" tube....... $
No. 80043—Nicoloi for 3" tube.......
No. 80042—Nicoloi for 2" tube.......

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".................... $
No. 80073—3"....................
No. 80072—2"....................

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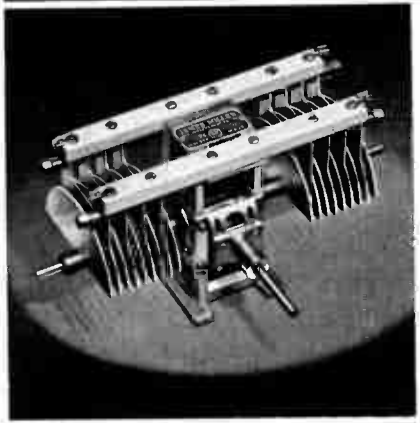
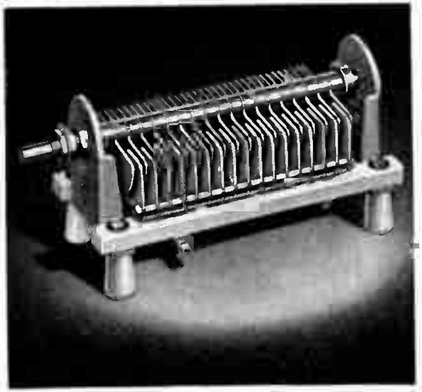
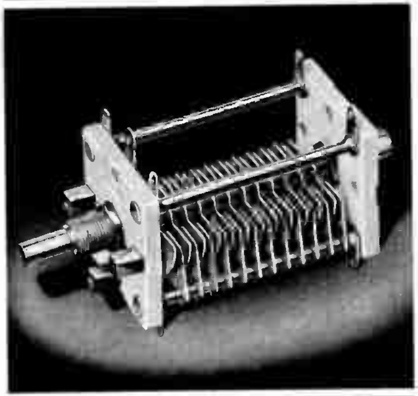
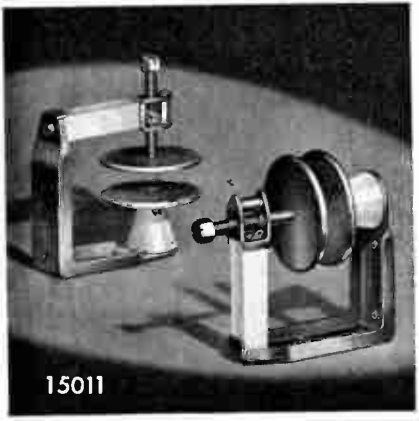








15011

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	$
11050	3000	50	
11070	3000	70	
04050	6000	50	
04060	9000	60	
04100	6000	90	
04200	3000	205	

12000 and 16000 SERIES TRANSMITTING CONDENSERS

Rigid heavy channeled aluminum end plates. Isolantite insulation, polished or plain edges. One piece rotor contact spring and 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 constant 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.

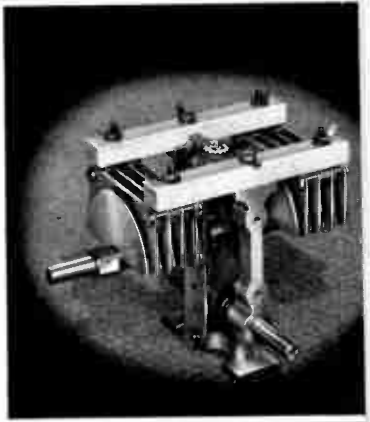
NEUTRALIZING CAPACITOR

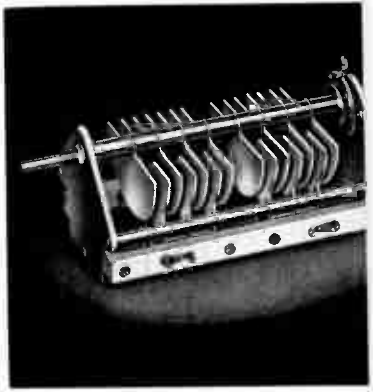
Designed originally for use in our own No. 90881 Power Amplifier, the No. 15011 disc neutralizing capacitor has such unique features as rigid channel frame, horizontal or vertical mounting, fine thread over-size lead screw with stop to prevent shorting and rotor lock. Heavy rounded-edged polished aluminum plates are 2" diameter. Glazed Steatite insulation.

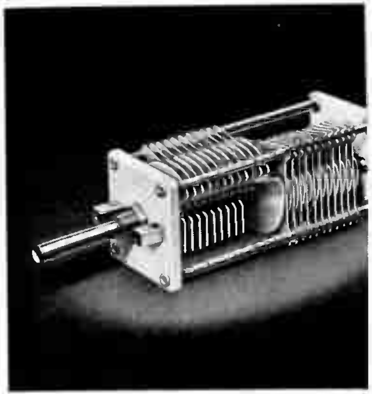
No. 15011 .... $

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.







68161

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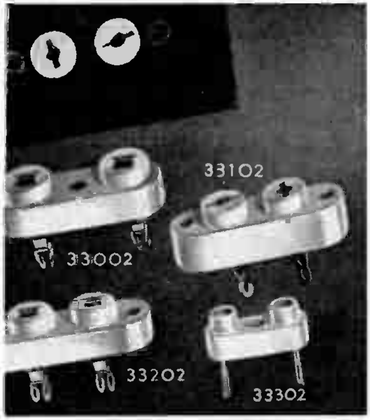
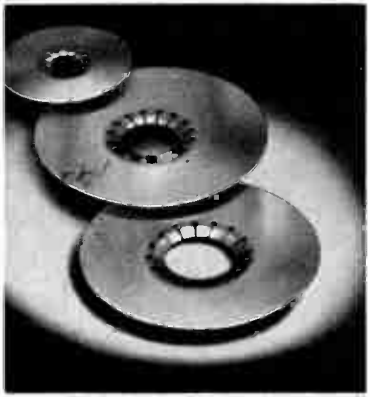
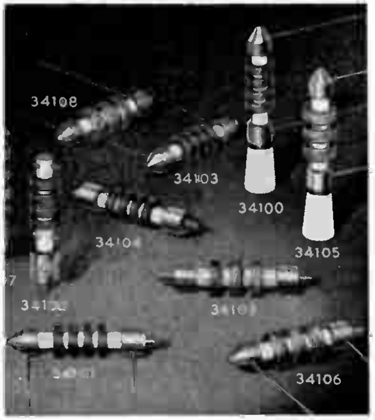






TUBE SOCKETS DESIGNED FOR APPLICATION

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, commerciol 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 ond crystal sockets also have barriers between individual contocts in addition.

The No. 33888 shield is for use with the 33008 octal socket. By its use, the electrostatic isolation of the grid ond plate circuits of single-ended metal tubes can be increased to secure greoter stobility 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 Contoct 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	$
No. 33005	
No. 33006	
No. 33007	
No. 33008	
No. 33888	
No. 33087	
No. 33002	
No. 33102	
No. 33202	
No. 33302	
No. 33446*	
No. 33991	
No. 33992	

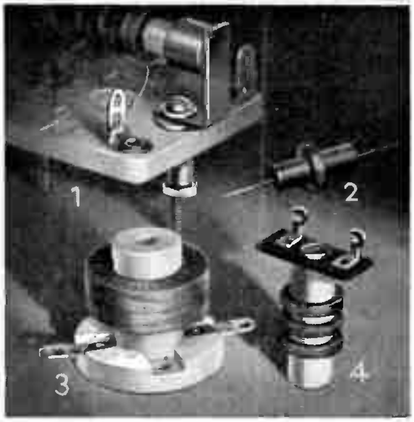
* For set of 3. Single discs $2.00 each.

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	$
No. 34101	
No. 34102	
No. 34103	
No. 34104	



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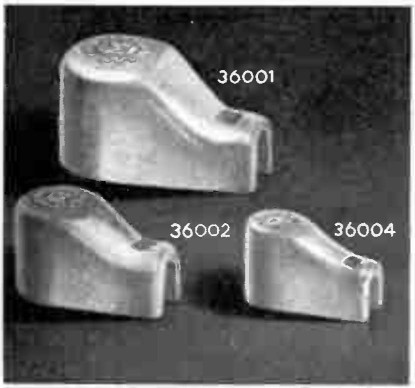


CERAMIC PLATE OR GRID CAPS

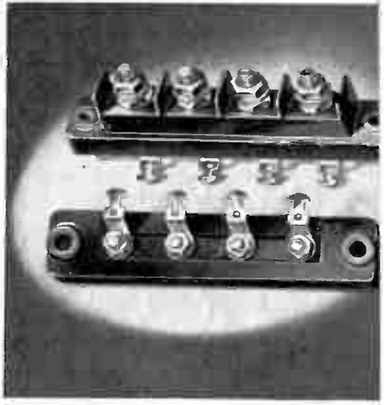
Soldering lug and contact one-piece. Lug ears annealed and solder dipped to facilitate easy combination "mechanical plus soldered" connection of cable.

No. 36001—9/16″............ $
No. 36002—⅜″...............
No. 36004—¼″...............

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″............ $
No. 36012—⅜″...............



SAFETY TERMINAL

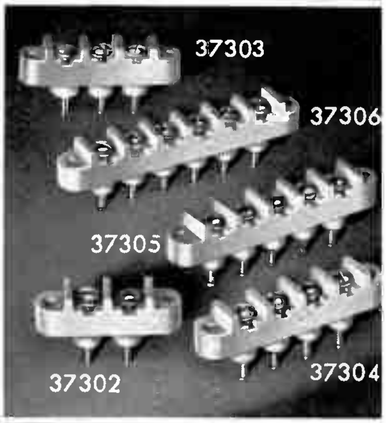
Combination high voltage terminal and thru-bushing. Tapered contact pin fits firmly into conical socket providing large area, low resistance connection. Pin is swivel mounted in cap to prevent twisting of lead wire.

No. 37001, Black or Red......... $
No. 37501, Low loss.............

TERMINAL STRIP

A sturdy four-terminal strip of molded black Textolite. Barriers between contacts. "Non turning" studs, threaded 8/32 each end.

No. 37104.................... $



POSTS, PLATES and PLUGS

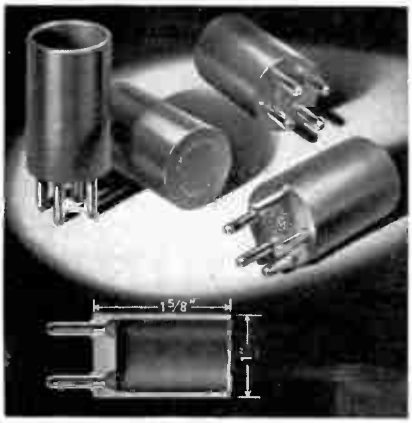
Designed for Application! Compact, easy to use. Made in black and red regular bakelite as well as low loss brown mica filled bakelite or steatite for R.F. uses. Posts have captive head.

No. 37202 Plates (pr.)........... $
No. 37212 Plugs................
No. 37222 Posts (pr.)............

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.................... $
No. 37303....................
No. 37304....................
No. 37305....................
No. 37306....................



MIDGET COIL FORMS

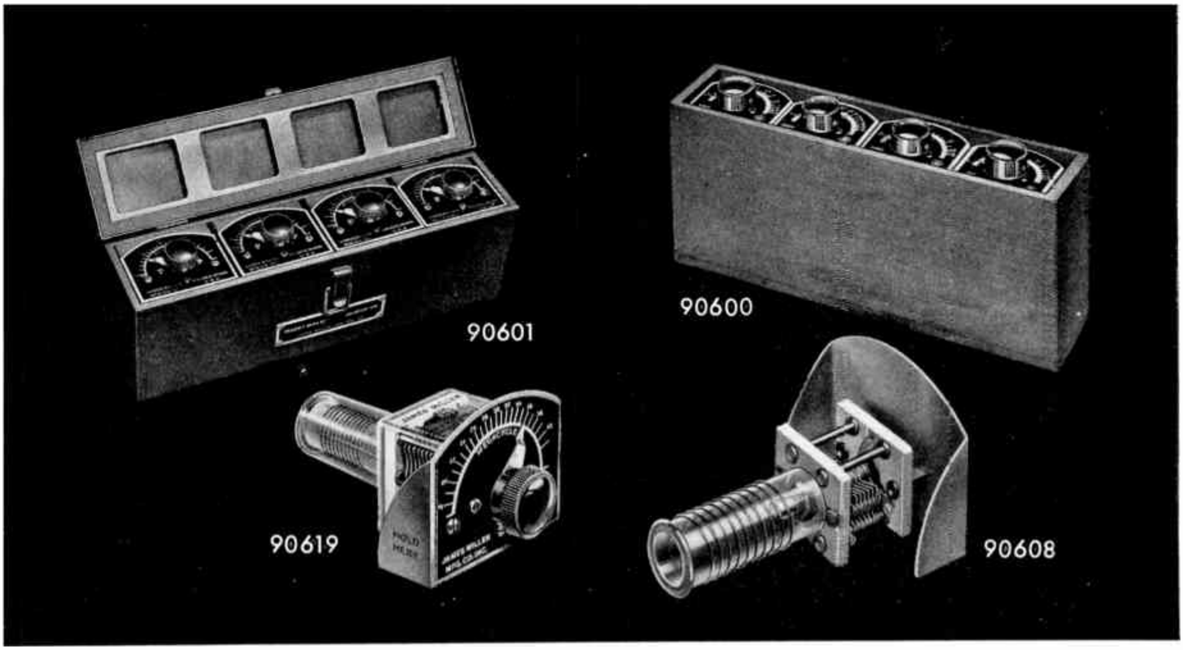
Made of low loss mica filled brown bakelite. Guide funnel makes for easy threading of leads through pins.

No. 45000.................... $
No. 45004....................
No. 45005....................

TUNABLE COIL FORM

Standard octal base of low loss mica-filled bakelite, polystyrene ½″ 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........ $
No. 74002, less iron core........



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.	$
90605	Range 3.0 to 10 mc.	
90606	Range 9.0 to 23 mc.	
90607	Range 23 to 60 mc.	
90608	Range 50 to 140 mc.	
90609	Range 130 to 170 mc.	
90610	Range 105 to 150 mc.	
90619	Range 350 to 1000 kc.—Neon Indicator	
90620	Range 150 to 350 kc.—Neon Indicator	
90625	Range 2 to 6 mc.—Neon Indicator	
90626	Range 5.5 to 15 mc.—Neon Indicator	
90600	Complete set of 90605 thru 90608, in case	
90601	Complete set Field type Frequency Meters in metal carrying case 1.5 to 40 mc.	

DISTRICT SALES OFFICES

NEW YORK
Cooper-DiBlasi
159 W. 14th Street

PHILADELPHIA
L. D. Lowery
1343 Arch Street

SEATTLE
V. Jensen
2616 Second Avenue

INDIANAPOLIS
V. MacNabb
915 Westfield Blvd.

SAN FRANCISCO
Russ Hines
234 Ninth Street

CHICAGO
G. G. Ryan
549 W. Washington Blvd.

BOSTON
Harry Gerber
739 Boylston Street

DALLAS
Jack Yount
1423 Pleasant Grove Dr.

LOS ANGELES
W. Bert Knight
10373 W. Pico Blvd.

KANSAS CITY
J. O. Schmitz
34th & Broadway Aves.

GREENSBORO, N. C.
Harrell, Atcheson & Adams
P.O. Box 2158

MINNEAPOLIS
Elliott Equipment Co.
712 6th Avenue South

CANADA
H. R. Gray
44 Danforth Road
Toronto

JAMES MILLEN MFG. CO., INC.
MAIN OFFICE AND FACTORY
150 EXCHANGE ST., MALDEN, MASSACHUSETTS, U. S. A.

E. F. JOHNSON COMPANY

WASECA, MINNESOTA

The products included on these six pages represent only part of the extensive JOHNSON line. Ask your distributor or write JOHNSON for complete catalog.

THE VIKING 1 TRANSMITTER KIT

Conservatively rated at 100 watts AM phone output, 115 watts CW. Incorpora features such as band-switching, crystal control or optional VFO input, pi-network out tuning and complete coverage of all amateur bands from 160 to 10 meters. In addit to amateur use, the Viking 1 is also designed to operate at frequencies assigned to mc commercial services.

VFO drive requirements are very slight. Only six volts of 7.5 mc. RF is required for output at 30 mcs., less for the 14 and 7 mc. bands. Two volts of 1.75 mc. VFO outpu ample excitation for 1.75 and 3.5 mc. output.

Delivering full output on phone with 115 volts 50/60 cycle line voltage, the tra mitter's power consumption is 350 watts. With line voltage between 105 and 120 vc performance is satisfactory.

In addition to being a completely self-contained, compact, and efficient 100 w transmitter, the Viking 1 can be used as a driver for a kilowatt amplifier. Full output the modulator is available at a nominal 500 ohms impedance.

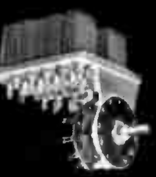
Clear, complete, easy-to-follow instructions make assembly easy—assure perf performance. Everything needed is included. No holes to drill, every part is furnish including cabinet, wiring harness, screws, nuts, washers, solder terminals, wire, gromm —everything!

See it at your jobbers today—or write for literature.

Tube Line Up

6AU6 crystal oscillator	6AU6 voltage amplifier	5R4 HV rectifie
6AQ5 buffer/doubler	6AU6 driver	5Z4 LV rectifier
4D32 final amplifier	807 pp modulators	6AL5 bias rectif

Complete, less tubes, crystals, key, mike Amateur Net **$209.**

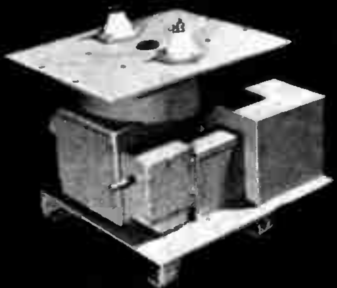


CRYSTAL SELECTOR

Ten frequencies with a twist of the knob with extra position for ECO. Accommoda crystals with ½″ spacing. With adaptors also takes ¾″ spaced holders. Bracket pern vertical or horizontal mounting.

126-220-1. Instant Crystal Selector.

126-120-1. Crystal Mounting Board only.



UNIVERSAL ROTOMATIC ANTENNA

Universal because its construction allows a variety of different combinations and typ of beams. Main boom alloy steel tubing to which elements are attached with spec clamps allowing any spacing or combination of elements.

Rotomatic—the deluxe Rotator. Simple to erect—built to last a lifetime. Heavy ov size gears, bearings, shafts. Precision throughout. Remote control box with selsyn directi indicator. Weatherproof RF relay box for switching with dual beams.

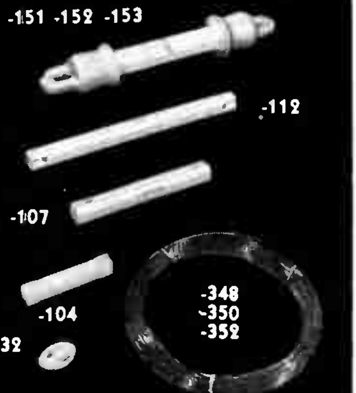
Fully adjustable aluminum alloy parasitic elements, heavier walls and larger diamete to withstand high winds and ice loading. High gain and front to back ratio.

Many combinations of single or dual arrays are available up to 4 elements on meters, 3 elements on 20 meters. Write for Catalog 204 for complete information.

Cat. No.

138-111. Rotator complete with motor & control box.

138-108. Weatherproof antenna relay for dual beams.



ANTENNA ACCESSORIES

Enamelled Copperweld Antenna Wire

Will not stretch nor sag. Carried by most suppliers in bulk, it is available from t factory in any specified length.

Cat. No.	B & S Gauge	Ft. per lb.	Break Streng
144-348	10	34½	1130 l
144-350	12	54	720 l
144-352	14	85	400 l

Feeder Insulators

136-122—2″ long **136-124**—4″ long **136-126**—6″ lc

Antenna Insulators

Cat. No.	Dia.	Break. Strng.	Lg.
136-104	⅝″ sq.	400 lbs.	4″
136-107	1″	800 lbs.	7″
136-112	1″	800 lbs.	12″

Commercial type 1½″ wet process porcelain, 5000 lbs. breaking strength.

Cat. No.	Length Net	Length Overall	Cat. No.	Length Net	Length Overall	Cat. No.	Length Net	Length Overa
136-151	8″	15½″	**136-152**	12″	19½″	**136-153**	20″	25½

Cat. No. **136-32** 1½″ Airplane type porcelain strain insulator.

. . a famous name in Radio

VARIABLE CONDENSERS

Partial Listing

his is a partial listing of the large JOHNSON line of quality condensers. Several types not shown, likewise many additional sizes are available in most types. All types ploy excellent steatite insulation. Approximate flashover voltage is 100x final nums in catalog numbers, (except Type N). "L" dimension is overall length less shaft ension.

TYPES C and D

dy, rigid construction at low cost! Aluminum plates .051 thick, rounded edges. Panel space e C, 5½" wide x 5⅜" high, Type D, 4¼" wide and 4" high.

TYPE C—Single Section

. No.	Max. Cap.	Air Gap	Number Plates	L
0C70	.252	.175"	24	6 13/16
0C70	.496	.175"	47	12 3/16
0C90	.337	.250"	43	14 27/32
C110	51	.350"	8	4 25/32
0C130	.102	.500"	21	13 11/32

TYPE C—Dual Section

Cat. No.	Max. Cap. Per Sec.	Air Gap	Number Plates	L
300CD70	.305	.175"	29	16 25/32
150CD90	.147	.250"	19	14 27/32
50CD110	50	.350"	8	10 5/16
100CD110	.103	.350"	17	16 25/32
50CD130	51	.500"	10	14 27/32

TYPE D—Single Section

0D35	.496	.080"	39	6 25/32
0D45	.146	.125"	17	4 25/32
D70	72	.175"	11	4 25/32
0D70	98	.175"	15	4 25/32
0D70	.244	.175"	37	10 5/16
0D90	99	.250"	19	7 11/16
0D90	.149	.250"	29	10 5/16

TYPE D—Dual Section

500DD35	.496	.080"	39	13 11/32
150DD45	.155	.125"	18	9 15/32
50DD70	52	.175"	8	5 13/16
70DD70	72	.175"	11	7 11/16
100DD70	97	.175"	15	9 15/32
150DD70	.151	.175"	23	13 11/32
50DD90	52	.250"	10	9 15/32

TYPES E and F

lugged compact units for low and medium power transmitters. Aluminum plates .032 thick, rounded ges. Stainless steel shafts. Panel space, Type E, 2⅝" square, Type F, 2 1/16" square.

TYPE E—Single Section

. No.	Max. Cap.	Air Gap	Number Plates	L
OE20	.353	.045"	33	3 17/32
OE20	.488	.045"	45	4 15/32
OE30	.100	.075"	15	2 9/16
OE30	.251	.075"	37	4 15/16
E45	53	.125"	12	2 31/32
OE45	145	.125"	33	6 3/32

TYPE E—Dual Section

Cat. No.	Max. Cap. Per Sec.	Air Gap	Number Plates	L
300ED20	.312	.045"	29	6 21/32
100ED30	99	.075"	15	5⅜
150ED30	.153	.075"	23	7 1/16
200ED30	.196	.075"	29	8⅜
50ED45	52	.125"	12	6 5/32
100ED45	.100	.125"	23	9 9/32

TYPE F—Single Section

OF20	.106	.045"	17	2¼
OF20	.154	.045"	25	2⅞
OF30	99	.075"	25	3 19/32
OF30	.148	.075"	37	4⅞

TYPE F—Dual Section

100FD20	.104	.045"	17	4 23/32
150FD20	.153	.045"	25	6
70FD30	66	.075"	17	5 23/32
100FD30	99	.075"	25	7 7/32

TYPE M MINIATURE

Smallest ever built, yet tops in accuracy. Ideal for VHF, miniature test equipment, etc. Panel space ' x ¾". Air gap .017. Mounts in ¼" hole.

Single

Cat. No.	Capacity Max.	Capacity Min.
A11	5.1	1.5
A11	8.7	1.7
M11	14.6	2.1
M11	19.7	2.6

Differential

Cat. No.	Capacity Max.	Capacity Min.
6MA11	5.0	1.5
9MA11	8.6	1.8
15MA11	14.2	2.3
19MA11	19.6	2.7

Butterfly

Cat. No.	Capacity Max.	Capacity Min.
3MB11	3.3	1.7
5MB11	5.3	2.1
9MB11	8.5	2.7
11MB11	11.0	3.2

TYPE L

Ceramic soldered—no eyelets or rivets to loosen. All brass, soldered construction. "Bright alloy, ated. Ideal for rough service. Panel space 1⅜" square. Air gap .030"; also furnished in .020" 50" and .080". In addition to those listed, also available in Differential types.

Single End Plate

Cat. No.	Cap. per Sect. Max.	Cap. per Sect. Min.	Number Plates
L15	11	2.8	3
L15	27	3.5	7
L15	51	4.6	13
L15	75	5.7	19

Double End Plate

0L15	99	6.8	25
0L15	202	11.6	51

Dual Section

Cat. No.	Cap. per Sect. Max.	Cap. per Sect. Min.	Number Plates
25LD15	27	3.5	7
50LD15	51	4.6	13
100LD15	99	6.8	25

Butterfly

10LB15	10.5	2.8	5
25LB15	26	4.3	12
50LB15	51	6.5	23

TYPE N

Small mounting space requirements, extremely gh voltage rating and fine adjustment make ese neutralizing condensers ideal.

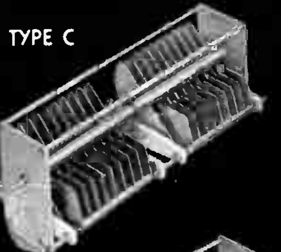
Cat. No.	Capacity Max.	Capacity Min.	Spacing
125	11.0	1.1	.125"
250	10.6	1.4	.250"
375	10.7	1.7	.375"

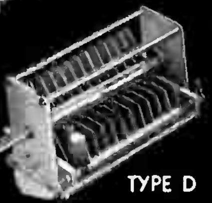
TYPE G

Extremely popular as neutralizing condensers for medium and low power stages. Also widely used for grid and plate tuning at high frequencies.

Cat. No.	Cap. per Sect. Max.	Cap. per Sect. Min.	Spacing
13G45	13	4.7	.125"
6G70	5.7	3.5	.225"
12G70	12	6	.225"

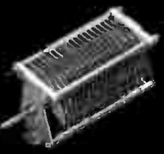
TYPE C

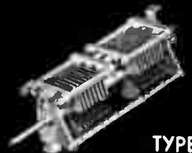




TYPE D

TYPE E





TYPE F

TYPE M

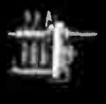
SINGLE

BUTTERFLY

DIFFERENTIAL

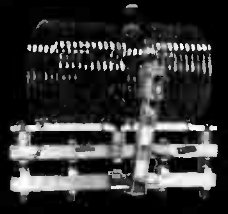
TYPE L SINGLE

TYPE L BUTTERFLY

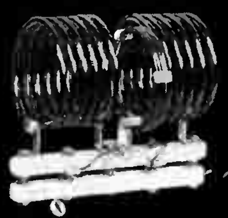


TYPE G

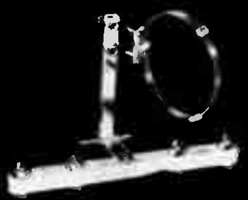
TYPE N



INDUCTOR 1000HCS40
LINK 1000SL5



INDUCTOR 500HCF20
LINK 150/500FL5



JACK BAR 1000JBS WITH
1000SLA ARM ASSEMBLY
& 1000SL5 LINK

JACK BARS
1000JBS, 500JBS, 150JBS

FARADAY SHIELD

114-220,
-260,-280

116-286

114-221,
-261,-281

114-262,-282

114-222

E. F. JOHNSON COMPANY

WASECA, MINNESOTA

AIR WOUND HAM INDUCTORS

JOHNSON Air Wound Ham Inductors provide a degree of e ciency never before available in commercially produced coils for amateur. This "broadcast" efficiency is possible because there model designed to match the impedance of each tank circuit—ei high voltage low current or low voltage high current tubes.

Efficiency is further increased because coil windings are a w size larger than on most available inductors—resulting in less heat lower loss with consequent higher efficiency.

JOHNSON Ham Inductors are built to give many years of efficient service. Coil winc are Formex-coated for better insulation and color preservation and JOHNSON qual apparent in the Steatite jack and plug bars and the crystal clear polystyrene coil sup and spacers. All JOHNSON inductors are conservatively rated.

The Swinging Link type inductors permit instant and perfect matching of inductor to transmission line thus preventing wasteful dissipation of power.

With these fine JOHNSON Ham Inductors and "plug-in" Swinging Link Assemblies, the amateur can instantly match coil to tube and link to line with broadcast efficiency. These outstanding inductors are also available in semi-fixed models.

HCS—Inductors match high voltage, low c rent tubes—swinging line type.

LCS—Inductors match low voltage, high c rent tubes—swinging link type.

HCF—Inductors match high voltage, low c rent tubes—semi-fixed link.

LCF—Inductors match low voltage, high c rent tubes—semi-fixed link.

Swinging Link

Cat. No.	Cat. No.	Cat. No.
1000HCS160	500LCS160	150LCS160
1000LCS160	500HCS80	150HCS80
1000HCS80	500LCS80	150LCS80
1000LCS80	500HCS40	150HCS40
1000HCS40	500LCS40	150LCS40
1000LCS40	500HCS20	150HCS20
1000HCS20	500LCS20	150LCS20
1000LCS20	500H/LCS14	150H/LCS14
1000H/LCS14	500H/LCS10	150H/LCS10
1000H/LCS10	500H/LCS6	150H/LCS6
500HCS160	150HCS160	

Semi-Fixed Link

Cat. No.	Cat. No.	Cat. No.
1000HCF80	500LCF80	150LCF80
1000LCF80	500HCF40	150HCF40
1000HCF40	500LCF40	150LCF40
1000LCF40	500HCF20	150HCF20
1000HCF20	500LCF20	150LCF20
1000LCF20	500H/LCF14	150H/LCF14
1000H/LCF14	500H/LCF10	150H/LCF10
1000H/LCF10	500H/LCF6	150H/LCF6
500HCF80	150HCF80	

Jack Bar Assemblies

Watts:	150	500	1000
Cat No.:	**150JBS**	**500JBS**	**100**

Swinging Link Arm Assemblie

150/500SLA. For 150/500 Watt Indu

1000SLA. For 1000 Watt Inductors.

Brackets

For Semi-Fixed Link Inductors.

150/500FLB. 150/500 Watt Bracket.

1000FLB. 1000 Watt Bracket.

"Plug-In" Links

Cat. No.	No. Turns	Cat. No.
150/500SL12	12	**1000SL10**
150/500SL5	5	**1000SL5**
150/500SL2	2	**1000SL2**

Semi-Fixed Links

Cat. No.	No. Turns	Cat. No.
150/500FL12	12	**1000FL10**
150/500FL5	5	**1000FL5**
150/500FL2	2	**1000FL2**

FARADAY SHIELD

Reduce TVI caused by capacitive coupling! Designed for JOHNSON Plug-In links, equally effective and easily installed on other links including non-plug-in types. Screen itself is metallic plating on polystyrene. Grounded hood and copper braid complete shielding. Link impedance relatively unchanged, plug-in link flexib unimpaired. Made in two sizes.

Cat. No.	Description
238-303.	150/500 watt swinging link sh hood and lead assembly.
238-304.	Same as above, for 1000 w
238-301.	150/500 watt link shield onl
238-302.	1000 watt link shield only.

NEW JOHNSON KNOBS & DIALS

Featuring fresh, advanced styling, these new JOHNSON models will enhance the pearance of your equipment. Molded phenolic knobs have 12 well defined flutes, lo gripping area. Knob faces slightly convex, sides slightly tapered to contribute to plea appearance. Beautiful satin chrome scales that will retain their new appearance ind nitely. Each knob and dial has brass set screw insert molded in place. Special mo available on quantity orders.

Knob Diam.	Shaft Diam.	Knob Only Cat. No.	Spinner Knob Cat. No.	Knob with Phenolic Skirt Cat. No.	Dia.	Knob with Chrome Dial Cat. No.	Diam.	Scale	
2⅜"	¼"	**116-280**	**116-286**	**116-281**	3"	**116-282**	4"	0-100	1
2⅜"	⅜"	**116-280-3**							
1⅝"	¼"	**116-260**		**116-261**	2¹⁄₁₆"	**116-262**	2¾"	0-100	1
1⅛"	¼"	**116-220**		**116-221**	1½"	**116-222-1**	1½"	100-0	1
1⅛"	¼"					**116-222-2**	1½"	0-10	2
1⅛"	¼"					**116-222-3**	1½"	1-7	1
1⅛"	¼"					**116-222-4**	1½"	On-off	
1⅛"	¼"					**116-222-5**	1½"	Indicator	

. . a famous name in Radio

TUBE SOCKETS

Highest Quality Sockets for Every Application

23-206. Industrial Bayonet, Steatite, Silver plated beryllium copper contacts. Base is 4 pin super jumbo. Tension springs in shell.
23-209. Medium 4 pin bayonet, white glazed porcelain base, metal shell, heavy phosphor bronze side wiping contacts. 2 13/16″ Dia.
23-209SB. Same as –209 but with Steatite base and beryllium copper contacts.
23-210. Same as –209 except contact to shell spacing not as great. 2½″ Dia.
23-211. Standard 50 watt type. Similar to –209 but with double filament contacts. 3⅜″ Dia.
23-211SB. Same as –211 but with Steatite base and beryllium copper contacts.
24-212. Steatite socket for RCA833 or 833A. 5⅛″ plate leads.
23-216. Giant 5 pin Bayonet. For tubes such as 803, RK28. 3¾″ Dia.
23-216SB. Same as –216 but with Steatite base and beryllium copper contacts.
24-213. For Eimac 152TL and 304TL. Contacts arranged for either series or parallel filaments.
24-214. For Eimac 1500TH, with air cooling jet.
24-215. For 250 watt tubes such as 204A, 849, etc. The plate terminal has a "safety cup" which prevents accidental dislodgement.

Wafer Types

Steatite, top and sides glazed. Brass contacts with steel springs cadmium plated.

22-217. 7 pin small.
122-224. 4 pin.
122-225. 5 pin.
122-226. 6 pin.
122-227. 7 pin medium.
122-228. Octal socket.
122-237. Giant 7 pin Steatite wafer. For Xmitting tubes such as HK257 and RCA813. With ¾″ diam. ventilating hole (not illustrated) in base.
122-247. 7 pin Steatite for tubes such as 826. Etched aluminum shield.
122-244. 4 pin Steatite. Super jumbo base tubes such as 8008.
122-101. 7 pin Steatite wafer with shield, retainer springs and provision for mounting button mica by-pass capacitors. Designed for VHF use with tubes such as 832.
122-275. Giant 5 pin Steatite wafer socket for 4-125A, RK48 tubes. Ventilation holes in base.

Miniature Sockets

120-267. all ceramic, 7 pin,
120-277B. with shield base, 7 pin.
133-277S. shield base only.

Shields

133-278A. 1⅜″ High, N.P. Brass
133-278B. 1¾″ High, " "
133-278C. 2¼″ High, " "

Acorn Type

121-265. Steatite acorn socket. Silver plated beryllium copper contacts.

COUPLINGS AND SHAFTS

JOHNSON insulated shaft couplings provide maximum voltage breakdown and superior strength. Glazed Steatite insulation except –264 which is phenolic.

Cat. No.	Mad. Peak Volt.	Dia.	Cat. No.	Mad. Peak Volt.	Dia.
104-250	4000	1 5/16″	**104-258**	————	½″
104-2503	4000	1 5/16″	**104-259**	8000	——
104-251	5000	2⅛″	**104-2593**	5000	——
104-251A	5000	2⅛″	**104-261**	7500	2½″
104-251B	5000	2⅛″	**104-262**	5000	2″
104-252	1000	11/16″	**104-264**	400	1½″

Panel Bearings

For ¼″ shaft. Up to ⅜″ panels.

115-255. Panel bearing only.
115-256. Bearing and 3″ shaft.
115-2562. Bearing and 6″ shaft.

Flexible Shafts

Non-rusting phosphor bronze, with ¼″ hubs, for connecting out of line control shafts.

115-253. 3″ long.
115-254. 6″ long.

PILOT, DIAL AND INDICATOR LIGHTS

JOHNSON dial and indicator light assemblies are outstanding examples of sound engineering design, excellent material and careful workmanship.

JOHNSON carries a complete line of hundreds of standard pilot light assemblies to meet every ordinary need. In addition, special assemblies can be furnished in production quantities on special order. Your inquiries are invited.


-206


-209, -210, -211, -216


-213


-215

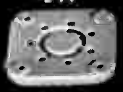

-212


-101


-247


-244


-275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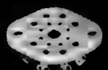
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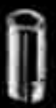
-277B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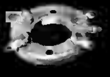
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-267


-261


-250


-251


-259
-2593


-252


-262


-264


-258


256
-2562


-253
-254

-45

-42

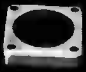
-42J

-20

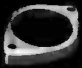
-50

-55

-22J



-866
-867



-865

-240
-241
-242

-54

-66

-53

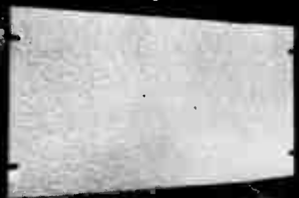
-503

-68

-15-1

-90

JOHNSON Aluminum Chassis are smoother, better fitted. No raw metal edges or openings at corners, nor overlapping metal. Spot-welded gussets on bottom flanges assure extreme rigidity. The one outstanding line!



Beautiful, fine black or silver gray wrinkle ⅛" aluminum panels. W.E. notching, 19" long, fit standard relay racks or cabinets.

E. F. JOHNSON COMPANY

WASECA, MINNESOTA

INSULATORS AND BUSHINGS

JOHNSON insulators are especially designed for high frequency use. They are m
of superior grade low absorption, well glazed electrical porcelain or Steatite. They
accurately molded and furnished with hardware of high grade nickel-plated brass.
JOHNSON insulators with confidence. "H" dimension is height of ceramic above panel

Stand-Off Insulators

STEATITE

Cat. No.	H	Hardware	Cat. No.	H	Hardw
135-20	1⁹⁄₁₆"	10-32	**135-22J**	1"	74-Ja
135-20J	1⁹⁄₁₆"	74-Jack	**135-24**	⅝"	6-32
135-22	1"	8-32			

PORCELAIN

135-60	4½"	¼-20	**135-62**	2¾"	¼-20

Metal Base Types

135-65	1⅜"	10-32	**135-67**	4½"	¼-20
135-65J	1⅜"	74-Jack	**135-67J**	4½"	76-Ja
135-66	2¾"	¼-20	**135-68**	2"	10-32
135-66J	2¾"	76-Jack	**135-68J**	2"	74-Ja

Steatite Cone Insulators

135-500	⅝"	6-32	**135-503**	2"	10-32
135-501	1"	8-32	**135-504**	3"	10-32
135-502	1½"	8-32			

Thru-Panel Insulators

STEATITE

135-40	1¼"	10-32	**135-42J**	⅞"	74-Ja
135-40J	1¼"	74-Jack	**135-44**	⅝"	6-32
135-42	⅞"	10-32			

PORCELAIN

135-45	1⅜"	10-32	**135-47**	4½"	¼-20
135-45J	1⅜"	74-Jack	**135-47J**	4½"	76-Ja
135-46	2¾"	¼-20	**135-48**	2"	10-32
135-46J	2¾"	76-Jack	**135-48J**	2"	74-Ja

Lead-In Bushings

STEATITE

135-50	½"	6-32	**135-52**	1⅛"	¼-20
135-51	¹³⁄₁₆"	10-32	**135-55**	¼"	6-32

PORCELAIN

135-53	1¾"	————	**135-54**	4"	———

Mounting flanges not included. See 135-90 and 135-91 below.

BOWL INSULATORS

Electrical glass, 6¹⁵⁄₁₆" OD, 4⅜" high. Fittings include ½" stud, nuts and washers, corona shield mounting flanges and gaskets.

Single Bowl	Two Bowls
135-15-0. Bowl only	**135-15-3.** 16" stud
135-15-1. 10¼" stud	**135-15-7.** 24" stud

MOUNTING FLANGES

Cat. No.
135-90 for bushing No. 135-53.

Cat. No.
135-91 for bushing No. 135-54.

THREADED BRASS ROD

¼" Dia. with nuts and washers.

Cat. No	Cat. No.	Cat. No.
115-240. 8" long	**115-241.** 10" long	**115-242.** 15" lo

ALUMINUM CHASSIS

Cat. No.	Size
18 GAUGE	
195-350	7x5x2
195-351	7x7x2
195-352	9x7x2
195-353	9½x5½x2
195-354	10x5x3
195-355	10x8x2½
195-356	11x7x2
195-357	12x7x3
195-358	12x8x2½
195-359	12x10x3
195-360	13x7x2
195-361	13x7x3
16 GAUGE	
195-363	10x14x3
195-364	15x7x3
195-366	17x4x3
195-367	17x7x3
195-368	17x8x2
195-369	17x8x3
195-370	17x10x2
195-371	17x10x3
195-372	17x10x4
195-373	17x10x5
195-374	17x11
195-375	17x11
14 GAUGE	
195-376	17x12
195-377	17x12
195-378	17x12
195-379	17x13
195-380	17x13
195-381	17x13
195-382	17x13

Also available in steel with permanent bright alloy plating.

ALUMINUM RACK PANELS

Black	Gray	Width	Black	Gray	Wid
196-161-4	**196-161-3**	1¾"	**196-167-4**	**196-167-3**	12¼
196-162-4	**196-162-3**	3½"	**196-168-4**	**196-168-3**	14"
196-163-4	**196-163-3**	5¼"	**196-169-4**	**196-169-3**	15¾
196-164-4	**196-164-3**	7"	**196-170-4**	**196-170-3**	17½
196-165-4	**196-165-3**	8¾"	**196-171-4**	**196-171-3**	19¼
196-166-4	**196-166-3**	10½"	**196-172-4**	**196-172-3**	21"

. a famous name in Radio

SPEED-X KEYS, PRACTICE SETS, BUZZERS

›tandard Semi-Automatic Keys

nproved model, heavy steel base, ber feet. Chrome plated vibrator and dware. Five adjustments, lowest and ıest speeds. Circuit closing switch. Ad-able paddles.

4-500. ⅛″ contacts, black wrinkle base.
4-501. ¼″ contacts, polished chrome base.
4-501L. Same as 114-501 except left-anded.

Amateur Special Model Semi-Automatic Key

ıam favorite, rubber feet, ⅛″ coin er contacts, chrome plated hardware ı vibrator, black wrinkle base.

4-515. Amateur model, semi-automatic.

Amateur Semi-Automatic Key With Switch

ʒimilar to Amateur Special but has :uit closing switch. Smaller, less weight.

4-510. Semi-Automatic with switch.

Heavy Duty Keys

Chrome plated key arm. ¼″ coin silver ıtacts. Navy knob.

4-320. Black wrinkle enamel base.
4-321. Polished chrome plated base.
4-326. Brass wrinkle finish base.

Standard Keys

High quality, low cost. Provision for plugging in semi-automatic key. ⅛″ coin silver contacts.

114-310. Black wrinkle, less switch.
114-3105. Black wrinkle, with switch.
114-311. Chrome plated, less switch.
114-3115. Chrome plated, with switch.
114-312. Gray wrinkle, less switch.
114-3125. Gray wrinkle, with switch.
114-316. Brass wrinkle, less switch.
114-3165. Brass wrinkle, with switch.

Molded Base Keys

Black phenolic base. ⅛″ coin silver contacts. Metal parts nickel plated.

114-301. Less switch.
114-3015. With switch.

Practice Keys

For beginners. ⅛″ coin silver contacts.

114-300. Molded brown phenolic base.
114-305. Black wrinkle finish metal base.

Practice Set

Constant frequency buzzer & key mounted on 4″ x 6″ phenolic base.

114-450. Code practice set.

Constant Frequency Buzzer

Fully adjustable, holds frequency. Uses 2 dry cells or "C" battery.

114-400. Constant frequency buzzer.

PLUGS AND JACKS

Banana Spring Type

Accurately turned from brass, with milled ›s and tinned terminals. Nickel plated. :kel-silver springs (other metals optional). ʌ contact resistance, high current ca-city.

–75 series plugs fit –74 series jacks, 7 series plugs fit –76 jack. –7451 and 452 have molded phenolic heads.

JACKS

ı8-74. ¼ — 28 x 17/32 thread.
8-7451. ¼ — 28 x ½ thread, red.
ı8-7452. ¼ — 28 x ½ thread, black.
ı8-76. ⅜ — 24 x 15/16 thread.

PLUGS

ı8-75. 6 — 32 x ⅜ thread.
ı8-75A. 6 — 32 x ¾ thread.
ı8-75BB. ⅜ x 1⅜ handle, black.
ı8-75BR. ⅜ x 1⅜ handle, red.
ı8-75C. 6 — 32 x 5/16 screw.
ı8-77. 10 — 32 x ⅜ thread.
ı8-77A. 10 — 32 x ¾ screw.
ı8-77BB. ⅝ x 1¾ handle, black.
ı8-77BR. ⅝ x 1¾ handle, red.

Tip Jacks and Plugs

PLASTIC HEAD TIP JACKS

Attractively colored strong Plaskon heads, accurately threaded ¼–32 with milled hex nut and insulating washers for ⅜ hole.

Cat. No.	Color	Cat. No.	Color
105-520	Red	**105-526**	Orange
105-521	Black	**105-527**	Yellow
105-522	Dk. Green	**105-528**	Lt. Green
105-524	Brown	**105-529**	Dk. Blue
105-525	Lt. Blue	**105-530**	Ivory

Molded Tip Jacks

Heavy duty type. Nickel plated brass body molded into phenolic head. 5/16–40 thread, and insulating washers for ⅜ hole.

No. 105-418. Red **No. 105-419.** Black

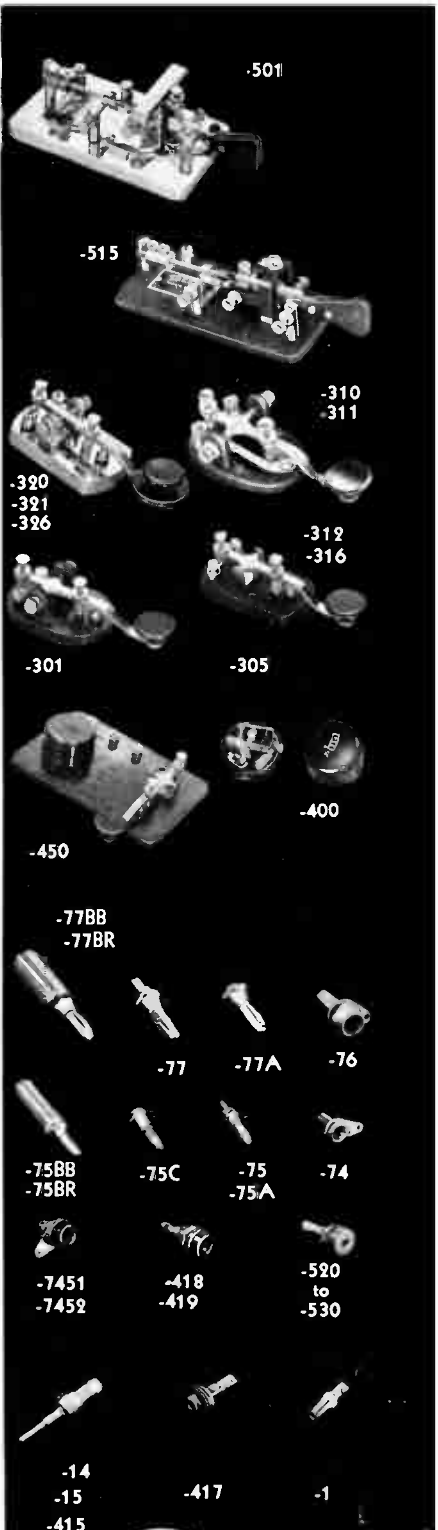
All Metal Tip Jack

Nickel plated brass, 5/16 hex head, ¼–32 thread, with insulating washers for ⅜ hole. 105-1 similar but headless, no nut nor washers, for mounting in ¼–32 tapped panel hole.

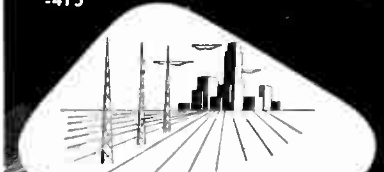
No. 105-417 **No. 105-1**

Solderless Tip Plugs

No. 105-15. 13/16 prong
No. 105-415. 9/16 prong
No. 105-14. Long, sharpened point



BROADCAST EQUIPMENT



Careful attention to specifications is the reason JOHNSON Broadcast Equipment enjoys :h wide popularity. Because practically all units are individually designed, there's no ed for compromises with good engineering. The cost is no more—frequently it is lower ın less flexible, less generously rated equipment. For specific information on the equip-:nt listed below and related items, write E. F. JOHNSON CO.

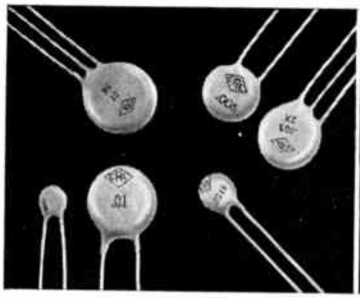
ıtenna Phasing Equipment
ıaxial Line
ılation Filters

Phase Sampling Loops
Pressurized Capacitors
Open Wire Line Supports

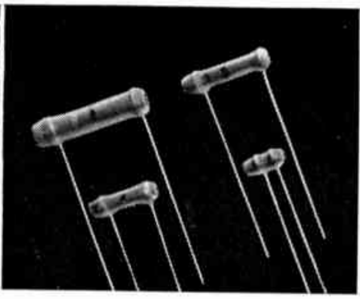
RF Contactors
Tower Lighting Filters
Variable Inductors

ARE YOU QUALITY-CONSCIOUS?

New Centralab parts give you the advantages of Centralab leadership in electronic component research

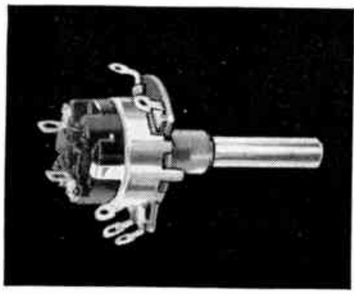


NEW! Centralab's new ceramic Hi-Kap capacitor line gives you discs and plates up to and including .1 mfd — tops in ceramics —your most permanent type capacitors.

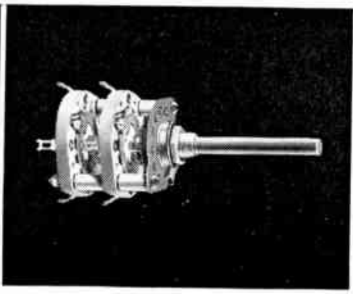


NEW! New BC Hi-Kap ceramic tubulars are now available in 48 different values. Best for r.f. by-pass and audio-coupling applications. For temperature compensating capacitors — ask for CRL's TC's.

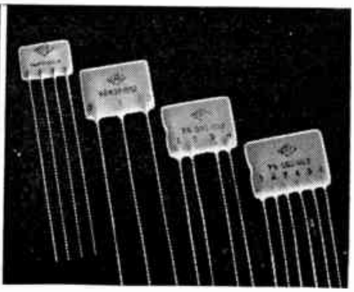
NEW! The best in ceramic stability, permanence and high-temperature characteristics in tubular form with ample external insulation. Pictured: CRL Cat. No. TV6-502 rated at .005 mfd. — 6000 V DC.



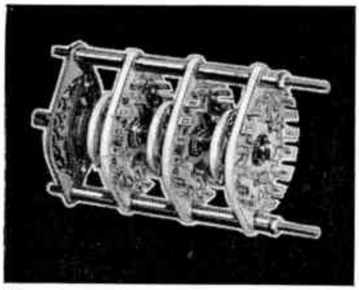
NEW! Centralab's new Blue Shaft Volume Controls are a complete line of all generally required sizes . . . plain and switch types. They're factory-assembled and tested . . . ready to install.



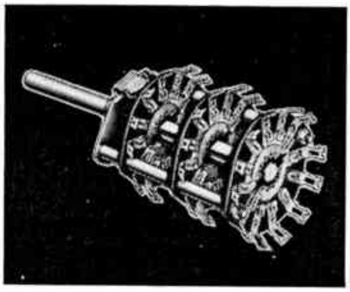
NEW! The famous CRL Ham Switch has a new look! It has heavier Steatite insulators — smoother action. You can't find a huskier amateur switch with a longer life . . . it's high quality at low cost.



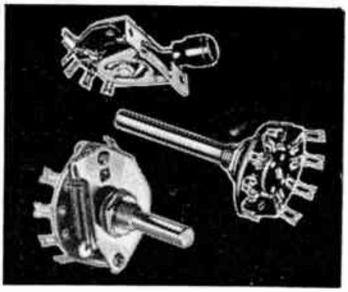
NEW! Now you can get CRL's famous Printed Electronic Circuits — everything from single value capacitors and resistor plates to complete 3-stage speech amplifiers. Ask your local CRL distributor.



POWER SWITCHES are specially designed for transmitters, power supply convertors and other applications. Efficient performance up to 20 megacycles.



ROTARY BAND SWITCH is used primarily for band change and general tap switch applications. Made with Steatite or phenolic insulation.



LEVER, SPRING RETURN, TONE SWITCHES. See your Centralab distributor for complete details on these switches — and all quality CRL parts.

The NEW products are listed in Catalog 27. Just tell us you are a radio amateur — it will be mailed at once.

Centralab CRL ®

Division of Globe-Union Inc., Milwaukee

You GET THE BEST IN *Heathkits*

Heathkits are the Quality Line of TEST INSTRUMENT KITS

Famous

HEATHKIT PARTS

- MALLORY FILTER CONDENSERS
- WILKOR PRECISION RESISTORS
- GRIGSBY ALLISON SWITCHES
- ALLEN-BRADLEY RESISTORS
- GENERAL ELECTRIC TUBES
- CHICAGO TRANSFORMER
- CENTRALAB CONTROLS
- SIMPSON METERS
- CINCH SOCKETS

Modern STYLING KITS THAT MATCH

Heathkits are styled in the most modern manner by leading industrial stylists. They add beauty and utility to any laboratory or service bench. There is a complete line of Heathkit instruments allowing a uniformity of appearance.

An attractive service shop builds a feeling of confidence. Many organizations have standardized on Heathkits providing uniform service departments.

There is no waste space or false effort to appear large in Heathkits — space on service benches is limited and the size of Heathkit instruments is kept as small as is consistent with good engineering practice.

Complete KITS WITH PARTS THAT FIT...

When you receive your Heathkit, you are assured of every necessary part for the proper operation of the instrument.

Beautiful cabinets, handles, two-color panels, all tubes, test leads where they are a necessary part of the instrument, quality rubber line cords and plugs, rubber feet for each instrument, all scales and dials ready printed and calibrated. Every Heathkit is 110 V 60 Cy. power tranformer operated by a husky transformer especially designed for the job. Heathkit chassis are precision punched for ease of assembly. Special engineering for simplicity of assembly is carefully considered.

Accuracy ASSURED BY PRECISION PARTS

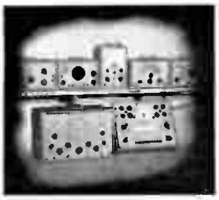
Wherever required, the finest quality 1% ceramic resistors are supplied. These require no aging and do not shift. No matching of common resistors is required. You find in Heathkit the same quality voltage divider resistors as in the most expensive equipment.

The transformers are designed especially for the Heathkit unit. The scope transformer has two electrostatic shields to prevent interaction of AC fields.

These transformers are built by several of the finest transformer companies in the United States.

Complete INSTRUCTION MANUALS

Heathkit instruction manuals contain complete assembly data arranged in a step-by-step manner. There are pictorials of each phase of the assembly drawn by competent artists with detail allowing the actual identification of parts. Where necessary, a separate section is devoted to the use of the instrument. Actual photos are included to aid in the proper location of wiring.



Used BY LEADING MANUFACTURERS

Leading TV and radio manufacturers use hundreds of Heathkits on the assembly lines. Heathkit scopes are used in the alignment of TV tuners. Impedance bridges are serving every day in the manufacture of transformers. Heathkit VTVM's are built into the production lines and test benches. Many manufacturers assemble Heathkits in quantity for their own use thus keeping purchase cost down.

Used BY LEADING UNIVERSITIES

Heathkits are found in every leading university from Massachusetts to California. Students learn much more when they actually assemble the instrument they use. Technical schools often include Heathkits in their course and these become the property of the students. High schools, too, find that the purchase of inexpensive Heathkits allows their budget to go much further and provides much more complete laboratories.

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The HEATH COMPANY
...BENTON HARBOR 26, MICHIGAN

12 Improvements IN NEW 1951

MODEL O-6 PUSH-PULL

Heathkit
OSCILLOSCOPE KIT

Only **$39.50**

- ★ New AC and DC push-pull amplifier.
- ★ New step attenuator frequency compensated input.
- ★ New non frequency discriminating input control.
- ★ New heavy duty power transformer has 68% less magnetic field.
- ★ New filter condenser has separate vertical and horizontal sections.
- ★ New intensity circuit gives greater brilliance.
- ★ Improved amplifiers for better response useful to 2 megacycles.
- ★ High gain amplifiers .04 Volts RMS per inch deflection.
- ★ Improved Allegheny Ludlum magnetic metal CR tube shield.
- ★ New synchronization circuit works with either positive or negative peaks of signal.
- ★ New extended range sweep circuit 15 cycles to over 100,000 cycles.
- ★ Both vertical and horizontal amplifier use push-pull pentodes for maximum gain.

The new 1951 Heathkit Push-Pull Oscilloscope Kit is again the best buy. No other kit offers half the features — check them.

Measure either AC or DC on this new scope — the first oscilloscope under $100.00 with a DC amplifier.

The vertical amplifier has frequency compensated step attenuator input into a cathode follower stage. The gain control is of the non frequency discriminating type — accurate response at any setting. A push-pull pentode stage feeds the C.R. tube. New type positioning control has wide range for observing any portion of the trace.

The horizontal amplifiers are direct coupled to the C.R. tube and may be used as either AC or DC amplifiers. Separate binding posts are provided for AC or DC.

The multivibrator type sweep generator has new frequency compensation for the high range it covers; 15 cycles to cover 100,000 cycles.

The new model 0-6 Scope uses 10 tubes in all — several more than any other. Only Heathkit Scopes have all the features.

New husky heavy duty power transformer has 50% more laminations. It runs cool and has the lowest possible magnetic field. A complete electrostatic shield covers primary and other necessary windings and has lead brought out for proper grounding.

The new filter condenser has separate filters for the vertical and horizontal screen grids and prevents interaction between them.

An improved intensity circuit provides almost double previous brilliance and better intensity modulation.

A new synchronization circuit allows the trace to be synchronized with either the positive or negative pulse, an important feature in observing the complex pulses encountered in television servicing.

The magnetic alloy shield supplied for the C.R. tube is of new design and uses a special metal developed by Allegheny Ludlum for such applications.

The Heathkit scope cabinet is of aluminum alloy for lightness of portability.

The kit is complete, all tubes, cabinet, transformer, controls, grid screen, tube shield, etc. The instruction manual has complete step-by-step assembly and pictorials of every section. Compare it with all others and you will buy a Heathkit. Model 0-6. Shipping Wt., 30 lbs.

New INEXPENSIVE MODEL S-2
ELECTRONIC SWITCH KIT

Twice as much fun with your oscilloscope — observe two traces at once — see both the input and output traces of an amplifier, and amazingly you can control the size and position of each trace separately — superimpose them for comparison or separate for observation — no connections inside scope. All operation electronic, nothing mechanical — ideal for classroom demonstrations — checking for intermittents, etc. Distortion, phase shift and other defects show up instantly. Can be used with any type or make of oscilloscope. So inexpensive you can't afford to be without one.

Has individual gain controls, positioning control and coarse and fine switching rate controls — can also be used as square wave generator over limited range. 110 Volt transformer operated comes complete with tubes, cabinet and all parts. Occupies very little space beside the scope. Better get one. You'll enjoy it immensely. Model S-2. Shipping Wt., 11 lbs.

Only **$19.50**

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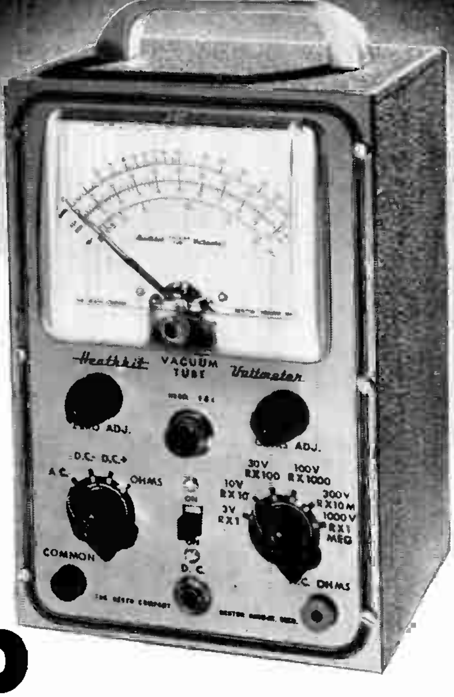
New 1951 • • MODEL V-4A

Heathkit

VTVM KIT

HAS EVERY EXPENSIVE *Feature*



★ Higher AC input impedance, (greater than 1 megohm at 1000 cycles).
★ New AC voltmeter flat within 1 db 20 cycles to 2 megacycles (600 ohm source).
★ New accessory probe (extra) extends DC range to 30,000 Volts.
★ New high quality Simpson 200 microampere meter.
★ New ½% voltage divider resistors (finest available).
★ 24 Complete ranges.
★ Low voltage range 3 Volts full scale (⅓ of scale per volt).
★ Crystal probe (extra) extends RF range to 250 megacycles.
★ Modern push-pull electronic voltmeter on both AC and DC.
★ Completely transformer operated isolated from line for safety.
★ Largest scale available on streamline 4½ inch meter.
★ Burn-out proof meter circuit.
★ Isolated probe for dynamic testing no circuit loading.
★ New simplified switches for easy assembly.

New LOW PRICE $23.50

The new Heathkit Model V-4A VTVM Kit measures to 30,000 Volts DC and 250 megacycles with accessory probes — think of it, all in one electronic instrument more useful than ever before. The AC voltmeter is so flat and extended in its response it eliminates the need for separate expensive AC VTVM's. + or − db from 20 cycles to 2 megacycles. Meter has decibel ranges for direct reading. New zero center on meter scale for quick FM alignment.

There are six complete ranges for each function. Four functions give total of 24 ranges. The 3 Volt range allows 33⅓% of the scale for reading one volt as against only 20% of the scale on 5 Volt types.

The ranges decade for quick reading.

New ½% ceramic precision are the most accurate commercial resistors available — you find the same make and quality in the finest laboratory equipment selling for thousands of dollars. The entire voltage divider decade uses these ½% resistors.

New 200 microampere 4½" streamline meter with Simpson quality movement. Five times as sensitive as commonly used 1 MA meters.

Shatterproof plastic meter face for maximum protection.

Both AC and DC voltmeter use push-pull electronic voltmeter circuit with *burn-out* proof meter circuit.

Electronic ohmmeter circuit measures resistance over the amazing range of 1/10 ohm to one billion ohms all with internal 3 Volt battery. Ohmmeter batteries mount on the chassis in snap-in mounting for easy replacement.

Voltage ranges are full scale 3 Volts, 10 Volts, 30 Volts, 100 Volts, 300 Volts, 1000 Volts. Complete decading coverage without gaps.

The DC probe is isolated for dynamic measurements. Negligible circuit loading. Gets the accurate reading without disturbing the operation of the instrument under test. Kit comes complete, cabinet, transformer, Simpson meter, test leads, complete assembly and instruction manual. Compare it with all others and you will buy a Heathkit. Model V-4A. Shipping Wt. 8 lbs. Note new low price $23.50

New 30,000 VOLT DC PROBE KIT

Beautiful new red and black plastic high voltage probe. Increases input resistance to 1100 megohms, reads 30,000 Volts on 300 Volt range. High input impedance for minimum loading of weak television voltages. Has large plastic insulator rings between handle and point for maximum safety. Comes complete with PL55 type plug.

No. 3366 High Voltage Probe Kit.
Shipping Wt., 2 pounds.

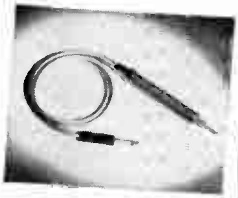
$5.50

Heathkit RF PROBE KIT

Crystal diode probe kit extends range to 250 megacycles ± 10% comes complete with all parts, crystal, cable and PL55 type plug.

No. 309 RF Probe Kit.
Shipping Wt., 1 lb.

$5.50



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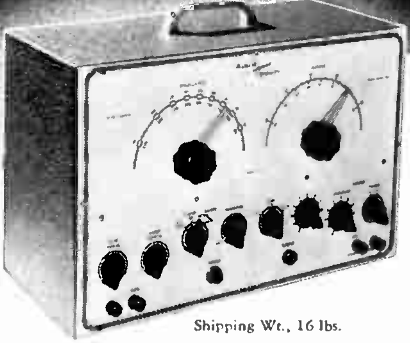
NEW *Heathkit* T. V. ALIGNMENT GENERATOR KIT



Shipping Wt., 16 lbs.

$39.50

- ★ New simplified circuit for easy calibration and assembly.
- ★ New 2 band built-in marker covers 19 to 75 Mc.
- ★ New dual spider sweep motor for long life.
- ★ New blanking circuit gives base line for better alignment.
- ★ New variable oscillator gives high output fundamentals on high TV band.
- ★ New standby switch keeps instrument ready for instant use.
- ★ New 6 to 1 slow speed drive on both master oscillator and marker tuners.

The new Heathkit TV Alignment Generator incorporates the new developments required fo modern TV servicing. An absorption marker circuit covering all possible IF bands and eve several of the RF bands. The new blanking circuit provides a base reference line which i invaluable in establishing proper traces. The new sweep motor incorporates dual spiders i the speaker frame assuring better alignment and long life. The mounting of the speake sweep motor has been simplified for easy alignment.

The variable master oscillator covers 140 to 230 Mc. thus giving high output fundamental where they are most needed. Low band coverage 2 Mc. to 90 Mc.

A new step attenuator provides excellent control of output.

Planetary 6 to 1 drives on both oscillator and marker provides smooth easy control settings

A standby position is provided making the instrument always instantly available.

Horizontal sweep voltage with phasing control is provided. No other sweep generator unde $100.00 provides all these features — comes complete with instruction manual Model TS 2

Heathkit CONDENSER CHECKER KIT

Only

$19.50

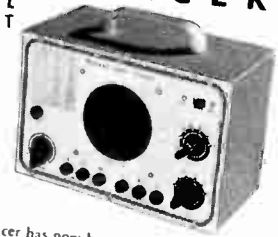
Features

- Power factor scale.
- Measures resistance.
- Measures leakage.
- Checks paper-mica-electrolytics.
- Bridge type circuit.
- Magic eye indicator.
- 110 V. transformer operated.
- All scales on panel.

Checks all types of condensers over a range of .00001 MFD to 1,000 MFD All on readable scales that are read direct from the panel. NO CHARTS OR MULTIPLIERS NECESSARY. A condeoser checker anyone can read. A leakage test and polarizing voltage for 20 to 500 Volts provided. Measures power factor of electrolytics between 0% and 50%. 110 V. 60 cycle transformer operated complete with rectifier and magic eye tube, cabinet, calibrated panel, test leads and all other parts. Clear detailed instructions for assembly and use. Model C-2. Shipping Wt., 7 lbs.

NEW *Heathkit* SIGNAL TRACER AND UNIVERSAL TEST SPEAKER KIT

$19.50



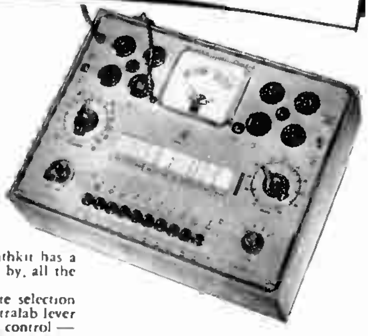
Features

- High sensitivity
- Complete set of speaker impedances
- Tests microphones and PA systems
- Tests both single and push-pull speaker circuits

The popular Heathkit Signal Tracer has now been combined with a universal test speaker at no increase in price. The same high quality tracer follows signal from antenna to speaker — locates intermittents — defective parts quicker — saves valuable service time — gives greater income per service hour. Works equally well on broadcast — FM or TV receivers. The test speaker has assortment of switching ranges to match push pull or single output impedance. Also test microphones, pickups, PA systems — comes complete — cabinet, 110 V. 60 cycle power transformer — tubes, test probe, all parts and detailed instructions for assembly and use Model T-2 Shipping Wt., 8 lbs.

Heathkit TUBE CHECKER KIT

Features



- Sockets for every modern tube — blank for new types.
- Fastest method of testing tubes — saves time — makes more profit.
- Rugged counter type birch cabinet.
- Gear driven roller chart gives instant setup for all types.
- Tests each element separately for open or short and quality.
- Beautiful 3 color meter — reads good-bad and line set point.

Test your tubes the modern way — dynamically — the simplest, yet fastest and surest method — your Heathkit has a switch for each tube element and measures that element — no chance for open or shorted elemeots slipping by, all the advantages of the mutual conductance type without the slow cumbersome time consuming setups.

Your Heathkit Tube Checker has all the features — beautiful 3 color BAD-GOOD meter — complete selection of voltages — roller chart listing hundreds of tubes including the new 9 pin miniatures — finest quality Centralab lever switches for each element — high grade birch counter type cabinet — continuously variable line adjust control — every feature you need to sell tubes properly The most modern type tube checker with complete protection against obsolescence The best of parts — rugged oversize 110 V. 60 cycle power transformer — finest of Mallory and Centralab switches and controls, complete set of sockets for all type tubes with blank spare for future types. Fast action brass gear driven roller chart quickly locates the settings for any type tube. Simplified switching cuts necessary testing time to minimum and saves valuable service time. Short and open element check Simple method allows instant setup of new tube types without waiting for factory data No matter what the arrangements of tube elements, the Heathkit flexible switching arrangement easily handles it Order your Heathkit Tube Checker Kit today See for yourself that Heath again saves you two-thirds and yet retains all the quality — this tube checker will pay for itself in a few weeks — better assemble it now Complete with instructions — pictorial diagrams — all parts — cabinet — ready to wire up and operate. Model TC-1 Shipping Wt. 12 lbs.

$29.50

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. . . BENTON HARBOR 26, MICHIGAN

NEW 1951 *Heathkit* SIGNAL GENERATOR KIT

Features

- **Sine wave audio modulation.**
- **Extended range 160 Kc. to 50 megacycles fundamentals.**
- **New step attenuator output.**
- **New miniature HF tubes.**
- **Transformer operated for safety.**
- **Calibrated harmonics to 150 megacycles.**
- **New external modulation switch.**
- **5 to 1 vernier tuning for accurate settings.**

A completely new Heathkit Signal Generator Kit. Dozens of improvements. The range on fundamentals has been extended to over 50 megacycles; makes this Heathkit ideal as a marker oscillator for T.V. New step attenuator gives controlled outputs from very low values to high output. A continuously variable control is used with each step. New miniature HF tubes are required for the high frequencies covered.

Uses 6C4 master oscillator and 6C4 sine wave audio oscillator. The set is transformer operated and a husky selenium rectifier is used in the power supply. The coils are precision wound and checked for calibration making only one adjustment necessary for all bands.

New sine wave audio oscillator provides internal modulation and is also available for external audio testing. Switch provided allows the oscillator to be modulated by an external audio oscillator for fidelity testing of receivers.

A best buy — think of all the features for less than $20.00. The entire coil and tuning assembly are assembled on a separate turret for quick assembly — comes complete — all tubes — cabinet — test leads — every part. The instruction manual has step-by-step instructions and pictorials. It's easy and fun to build a Heathkit Model SG-6 Signal Generator. Shipping Wt., 7 lbs.

$19.50

Heathkit SINE AND SQUARE WAVE AUDIO GENERATOR KIT

Either sine or square wave.
Stable RC bridge circuit.
Covers 20 to 20,000 cycles.
Less than 1% distortion.

Hundreds of Heathkit Audio Generators are used by speaker manufacturers—definite proof of their quality and dependability. The added feature of square wave opens up an entirely new field of amplifier testing. Uses the best of parts, 4 gang condenser, 1% calibrating resistors, metal cased filter condensers, 5 tubes, completely calibrated panel and detailed instruction manual. One of our best and most useful kits. Model G-2. Shipping Wt., 12 lbs.

$34.50

THE NEW *Heathkit* HANDITESTER KIT

- **Beautiful streamline Bakelite case.**
- **AC and DC ranges to 5,000 Volts.**
- **1% Precision ceramic resistors.**
- **Convenient thumb type adjust control.**
- **400 Microampere meter movement.**
- **Quality Bradley AC rectifier.**
- **Multiplying type ohms ranges.**
- **All the convenient ranges 10-30-300-1,000-5,000 Volts.**
- **Large quality 3" built-in meter.**

$13.50

A precision portable volt-ohm-milliammeter. An ideal instrument for students, radio service, experimenters, hobbyists, electricians, mechanics, etc. Rugged 400 ua meter movement. Twelve complete ranges, precision dividers for accuracy. Easily assembled from complete instructions and pictorial diagrams. An hour of assembly saves one-half the cost Order today. Model M-1 Shipping Wt., 2 lbs.

NEW *Heathkit* BATTERY ELIMINATOR KIT

Features

- **Provides variable DC voltage for all checks.**
- **Locates sticky vibrators-intermittents.**
- **Voltmeter for accurate check.**
- **Has 4000 MFD Mallory filter for ripple-free voltage.**

Even the smallest shop can afford the Heathkit Battery Eliminator Kit. A few auto radio repair jobs will pay for it. It's fast for service, the voltage can be lowered to find sticky vibrators or raised to ferret out intermittents. Provides variable DC voltage 5 to 7½ Volts at 10 Amperes continuous or 15 Amperes intermittent. Also serves as storage battery charger Ideal for all auto radio testing and demonstrating.

$22.50

A well filtered rugged power supply uses heavy duty selenium rectifier, choke input filter with 4,000 MFD of electrolytic filter for clean DC. 0-15 V voltmeter indicates output which is variable in eight steps Easily constructed in a few hours from our instructions and diagrams — better be equipped for all types of service — it means more income. Model BE-2. Shipping Wt., 19 lbs.

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The HEATH COMPANY
. . . BENTON HARBOR 26, MICHIGAN

New
LABORATORY
INSTRUMENT KITS

HUNDREDS OF LABORATORIES USE

Heathkit IMPEDANCE BRIDGE *as Standard*

Features

● Measures inductance 10 microhenries to 100 henries ● Measures resistance .01 ohms to 10 megohms ● Measures capacitance .00001 MFD to 100 MFD ● Measures "Q" and power factor.

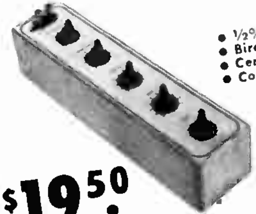
Measures inductance from 10 microhenries to 100 henries, capacitance from .00001 MFD to 100 MFD. Resistance from .01 ohms to 10 megohms. Dissipation factor from .001 to 1. "Q" from 1 to 1,000. Ideal for schools, laboratories, service shops, serious experimenters. An impedance bridge for everyone — the most useful instrument of all, which heretofore has been out of the price range of serious experimenters and service shops. Now at the lowest price possible. All highest quality parts. General Radio main calibrated control. General Radio 1,000 cycle hummer. Mallory ceramic switches with 60 degree indexing — 200 microamp type binding posts with standard ¾" centers. Beautiful birch cabinet. Directly calibrated "Q" and dissipation factor scales. Ready calibrated capacity and inductance standards of Silver Mica, accurate to ½ of 1% and with dissipation factors of less than 30 parts in one million. Provisions on panel for external generator and detector. Measure all your unknowns the way laboratories do — with a bridge for accuracy and speed.

$69.50

Internal 6 Volt battery for resistance and hummer operations. Circuit utilizes Wheatstone, Hay and Maxwell circuits for different measurements. Supplied complete with every quality part — all calibrations completed and instruction manual for assembly and use. Deliveries are limited. Model IB-1. Shipping Wt., 15 lbs.

NEW *Heathkit* LABORATORY RESISTANCE DECADE KIT

Features



- ½% Accuracy
- Birch Cabinet
- Ceramic Switches
- Covers 1 ohm to 99,999 ohms

The new Heathkit Resistance Decade is a handy tool for laboratory, school and service shop. Ideal for test setups, calibrating instruments, bridge measurements, selecting multipliers, etc.

$19.50

Uses the finest Centralab ceramic switches, ½% ceramic decade resistors and heavy birch cabinet matching other laboratory equipment. The range is 1 ohm to 99,999 ohms in one ohm steps.

Finest quality throughout to withstand school usage — heavy aluminum panel — laboratory type binding posts — the fine decades are extremely simple to assemble — complete kit. Model RD-1. Shipping Wt., 4 lbs.

NEW *Heathkit* LABORATORY POWER SUPPLY KIT

Features

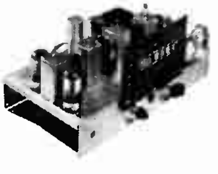
- Supplies 6.3 V. AC at 4.5 Amps.
- Heavy duty construction.
- Handy for schools, labs., and service shops.
- Supplies variable DC 50-300 Volts.
- Shows voltage or current on 3½" meter.

This new Heathkit Variable Power Supply Kit fills hundreds of needs — use it for experimental circuits — no need to build a separate power supply — use it for a test voltage to determine proper coefficients in unknown circuits — calibrate instruments with its variable voltage, etc. This new Heathkit supplies 50 to 300 Volts continuously variable DC together with an AC filament voltage of 6.3 Volts at 4.5 Amperes. A built-in 1 MA 3½" meter has proper shunts to read 0-500 Volts and 0-200 Milliamperes. The circuit uses a 5Y3 rectifier, two 1619 tubes as electronic control tubes to vary the output voltage with a single potentiometer. Case measures 7⅛" x 13" x 7⅛". Has instruction manual for assembly and use. Model PS-1. Shipping Wt., 18 lbs.

$29.50

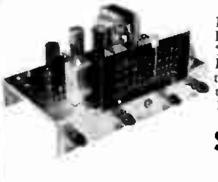
Heathkit RECEIVER & TUNER KITS for AM and FM

TWO HIGH QUALITY *Heathkit* SUPERHETERODYNE RECEIVER KITS



Model BR-1 Broadcast Model Kit covers 550 to 1600 Kc. Shipping Wt., 10 pounds.

$19.50



Model AR-1 3 Band Receiver Kit covers 550 Kc. to over 20 Mc. continuous. Extremely high sensitivity. Shipping Wt., 10 lbs.

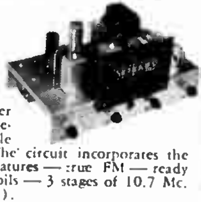
$23.50

Two new Heathkits. Ideal for schools, replacement of worn out receivers, amateurs and custom installations.

Both are transformer operated quality units. The best of materials are used throughout — six inch calibrated slide rule dial — quality power and output transformers — dual iron core shielded I.F. coils — metal filter condensers and all other parts. The chassis has phono input jack — 110 Volt outlet for phono motor and there is a phono-radio switch on panel. A large metal panel simplifying installation in used console cabinets is included. Comes complete with tubes and instruction manual incorporating pictorials and step-by-step instructions (less speaker and cabinet). The three band model has simple coil turret which is assembled separately for ease of construction.

TRUE FM FROM *Heathkit* FM TUNER KIT

$22.50



The Heathkit FM Tuner Model FM-2 was designed for best possible tonal reproduction. The circuit incorporates the most desirable FM features — true FM — ready wound and adjusted coils — 3 stages of 10.7 Mc. I.F. (including limiter).

Tube lineup: 7E5 oscillator, 6SH7 mixer, two 6SH7 I.F. stages, 6SH7 limiter, two 7C4 diodes as discriminator, 6X5 rectifier.

The instrument is transformer operated making it safe for connection to any type receiver or amplifier. The R.F. coils are ready wound — mounted on the tuning condenser and the condenser is adjusted — no R.F. coils to wind or adjust.

A calibrated six inch slide rule dial has vernier drive for easy tuning. The finest parts are provided with all tubes, punched and formed chassis, transformers, condensers and complete instruction manual. Model FM-2. Shipping Wt., 10 lbs.

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ENJOY MUSIC AT ITS *Finest* WITH *Heathkit* AMPLIFIERS

NEW *Heathkit* AMPLIFIER KIT

HIGH FIDELITY 20 WATT

Features

- Push-pull 6L6's.
- Full 20 Watts output.
- Fully enclosed chassis.
- Provisions for reluctance pickup compensation stage.
- Cased high fidelity output transformer.
- Treble and bass boost tone controls.
- Full range of output impedances 3.2 ohms to 500 ohms.

$21.50

The finest amplifier kit we have ever offered — check the features. This inexpensive amplifier compares favorably with instruments costing five times as much. Nothing has been spared to provide the best reproduction — an ideal amplier for the new Heathkit FM Tuner listed below.

Dual tone controls for control of both treble and bass. Bass control is of the boost type for maximum listening pleasure. Optional preamplifier stage for use with G. E. reluctance pickup or microphone. Uses inverse feedback to give excellent response over entire range. Tube lineup: 6SJ7 preamplifier stage, 6J5 phase splitter stage, two 6L6's in push-pull and 5Y3 rectifier. (6SC7 as optional compensation stage).

Uses highest quality Chicago Transformer Corporation cased output transformer with taps of 3.2, 8, 15, 60 and 500 ohms to match any speaker combination. Power transformer is conservatively rated for continuous operation in sound systems. Tone control gives maximum bass boost of 6 db at 70 cycles. Amplifier has maximum gain of 75 db. Response within 3 db 20 to 20,000 cycles. Shipping Wt., 17 lbs. Complete with all parts, tubes and instruction manual.

Model A-5A Amplifier with preamplifier for G. E. cartridges or microphone **$23.50**
12" 20 Watt Speaker, No. 326.. **7.50**

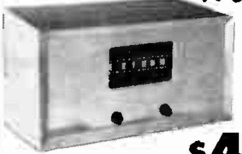
Heathkit ECONOMY 6 WATT PUSH-PULL AMPLIFIER KIT

$12.50

No. 304, 12-inch Speaker... **$6.95**

This new Heathkit Amplifier was designed to give quality reproduction at a very low price. Has two preamp stages, phase inverter stage and push-pull beam power output. Comes complete with six tubes, quality output transformer (to 3-4 ohm voice coil), husky cased power transformer and all other parts. Has tone and volume controls. Instruction manual has pictorial for easy assembly. Six watt output with response flat ± 1½ db from 50 to 15,000 cycles. A quality amplifier kit at new low price. Better build one. Model A-4. Shipping Wt., 7 lbs.

Heathkit RECEIVERS and TUNER CABINETS



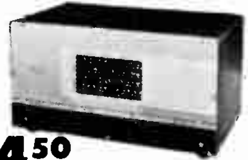
$4.95

Order No. 350 for FM tuner.

Blonde birch veneer cabinet for either the receivers or tuner. Modern styling is an asset to any room. 5" speaker fits in end of cabinet when used with receivers. Size 7 x 13½ x 8¼ inches. Shipping Wt., 5 lbs.
Order No. 345 for either receiver

Metal professional type communications receiver cabinet. Finished in deep grey to fit the panel supplied with Heathkit BR-1 and AR-1 Receivers (panel shown not included with cabinet). 5" speaker mounts in end of cabinet. Gives professional appearance to Heathkit receivers. Size 7 x 14 x 7¾ inches. Shipping Wt., 6 lbs.

5" Permoflux Speaker for either cabinet for use with either Heathkit Receiver
No. 320 5" Speaker.. **$2.75**



$4.50

No. 335 Cabinet for receivers only.

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Quantity	Item	Price	Quantity	Item	Price
	Heathkit Oscilloscope Kit — Model O-6			Heathkit VTVM Kit — Model V-4A	
	Heathkit T.V. Alignment Gen. Kit — TS-2			Heathkit R.F. Probe Kit — No. 309	
	Heathkit FM Tuner Kit — FM-2			Heathkit H.V. Probe Kit — No. 336	
	Heathkit Broadcast Receiver Kit — Model BR-1			Heathkit R.F. Signal Gen. Kit — Model SG-6	
	Heathkit Three Band Receiver Kit — Model AR-1			Heathkit Condenser Checker Kit — Model C-2	
	Heathkit Amplifier Kit — Model A-4			Heathkit Handitester Kit — Model M-1	
	Heathkit Amplifier Kit — Model A-5 (or A-5A)			Heathkit Variable Power Supply Kit — Model PS-1	
	Heathkit Tube Checker Kit — Model TC-1			Heathkit Resistance Decade Kit — Model RD-1	
	Heathkit Audio Generator Kit — Model G-2			Heathkit Impedance Bridge Kit — Model IB-1	
	Heathkit Battery Eliminator Kit — Model BE-2			Heathkit Signal Tracer Kit — Model T-2	
	Heathkit Electronic Switch Kit — Model S-2				

On Parcel Post Orders, include postage for weight shown and insurance. (We insure all shipments.)

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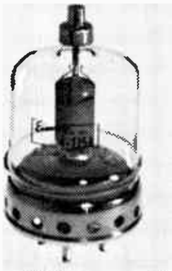
TUBES

THESE ARE THE TUBES OF CHAMPIONS . . . THE FIRST CHOICE OF LEADING AMATEURS EVERYWHERE. THEY PROVIDE MORE WATTS PER DOLLAR COST AND THE UTMOST IN DEPENDABILITY AND PERFORMANCE.

COMPLETE TUBE DATA AVAILABLE FREE

4-65A This small tetrode operates well at plate voltages from 600 to 3000 volts. At 2000 volts one tube will handle up to 300 watts input for CW or 240 watts for phone. Driving power is 2 to 3 watts.

4-125A The tube that made transmitting screen-grid tubes popular. The 4-125A will take a plate input of 500 watts for CW or 380 watts for phone. Driving power is less than 2 watts. A pair of these tubes makes an ideal high-power phone or CW final.



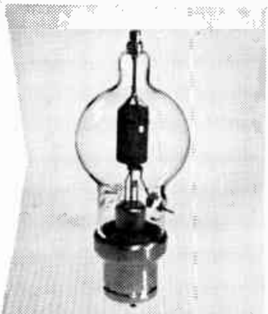
4-250A A pair of 4-250A tetrodes will easily handle a KW for phone. In CW service, one tube will take a KW input. Driving power is only 2 to 3 watts per tube. As modulators, a pair will deliver as much as 750 watts of audio with zero driving power.

4-400A For really de-luxe equipment use the 4-400A tetrode. One tube can be run at a KW input for CW or 880 watts for phone. Low drive, too, of course, less than 5 watts. Available as an accessory is an air-system socket for simplified cooling.

4X150A For VHF or UHF work, use the 4X150A. This small, forced-air cooled external anode tetrode will handle 250 watts input on the ultra-highs with a driving power of but a few watts. The 4X150A operates well on plate voltages as low as 400 or 500 volts, making it ideal for portable or mobile equipment with a wallop.

100T The "old reliable" 100T, improved throughout the years, is still top choice of amateurs who prefer a triode. Will take a plate input of 500 watts for CW or 350 watts for phone. As modulators, a pair will deliver 420 watts of audio.

250T A tried, proven, and continually improved 250-watt triode. The ideal triode for 1 KW CW input. Will handle 825 watts input on phone. With plate voltage as low as 1500 volts, a pair will modulate all the law allows.



VVC60-20 This is but one type in the Eimac line of variable and fixed vacuum capacitors for plate tank circuits. The VVC60-20 is variable over a range of 10 mmfd to 60 mmfd. Maximum r-f voltage is 20 kv.

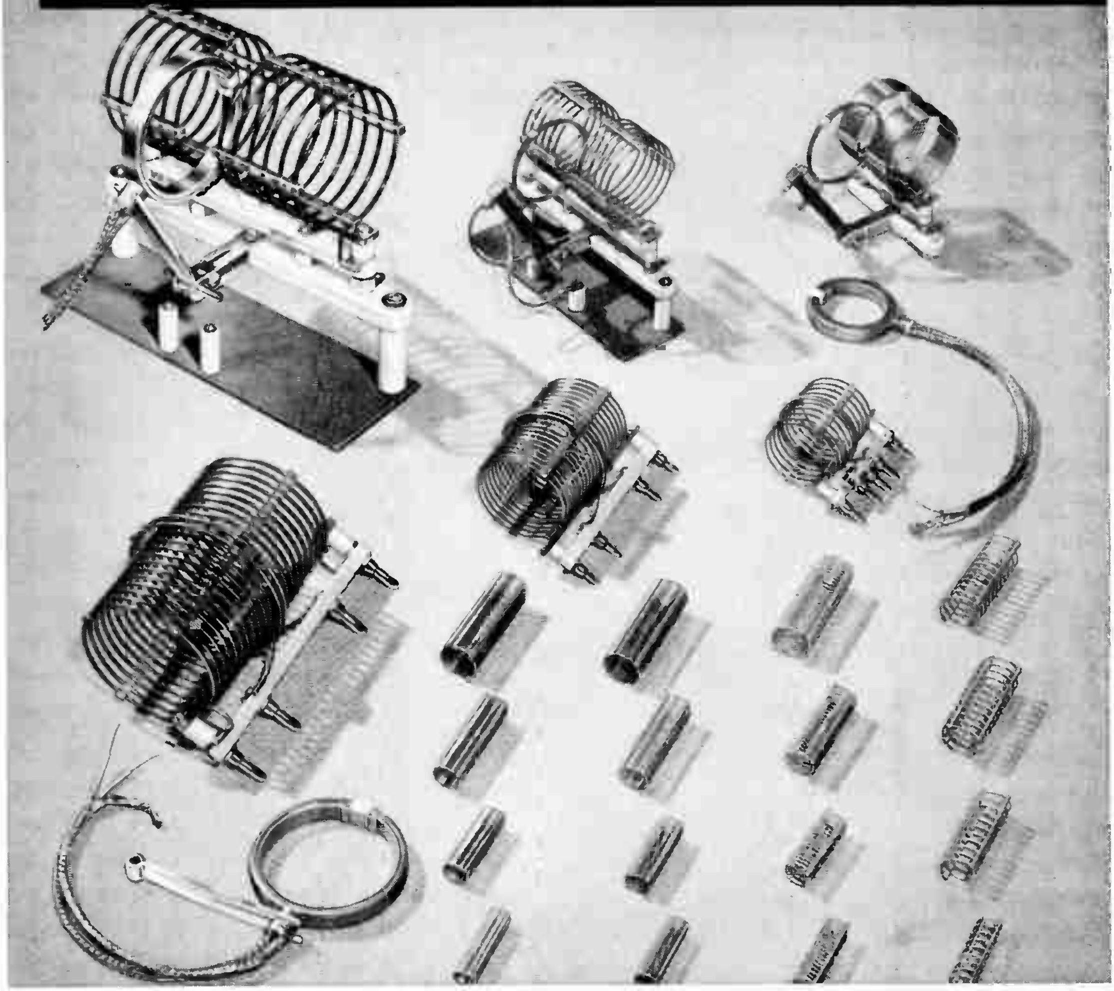
EITEL-McCULLOUGH, INC.

San Bruno, California

Export Agents: Frazar & Hansen, 301 Clay St., San Francisco, California

B&W AIR INDUCTORS

Pioneers of the Coil Industry



B&W was the first to develop and manufacture the air wound type inductor, a modern type of coil that sets a standard of design and construction throughout the industry.

Pioneers in many new types of coils, B&W was the first to manufacture a complete line of coils for amateur use. First in the development of variable links, B&W now offers "plug-in" variable links for greater flexibility.

Among the many other B&W firsts in the electronic field are: transmitting turret assemblies, coil-variable condenser combinations, miniature air wound coils and the latest development, the Faraday Shielded Link.

Sixteen years of unchallenged leadership provides an assurance that, regardless of the application, B&W has a coil to fit the need . . . and you can depend on them.

SEE NEXT PAGE

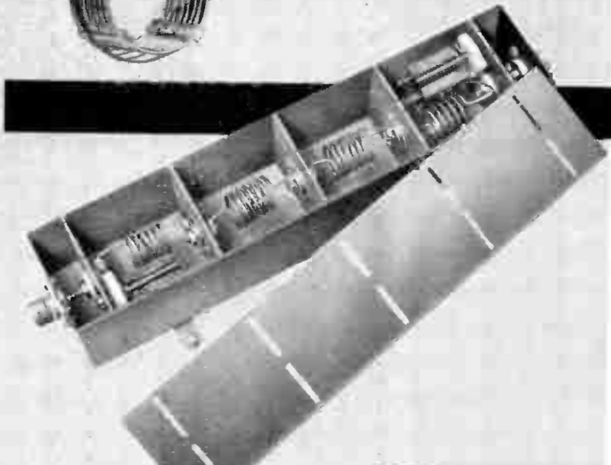
BARKER & WILLIAMSON, Inc.

237 Fairfield Ave., Upper Darby, Pa.

B&W PARTS an

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.



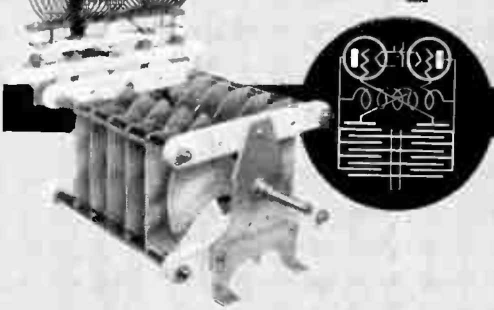
B&W LOW-PASS FILTERS

. . . Solve TV Interference Problems

Used in conjunction with the B&W Faraday Shielded Link, B&W Low-Pass Filters are highly effective in reducing Television interference caused by Amateur Transmissions. While the Faraday Link effectively reduces capacity coupled harmonic and spurious signal radiations, the Low-Pass Filter, consisting of two "M" derived end sections and three midsections of the constant "K" type, attenuate all frequencies above 30 megacycles 75 db or more throughout the entire TV band. Write for a copy of "Filter-Facts" priced at 15 cents for details on how to lick your TVI.

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

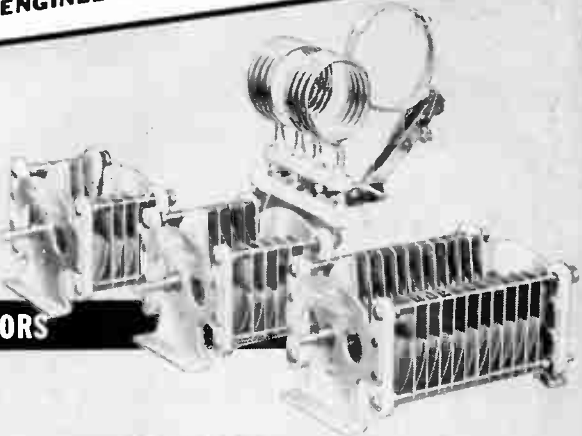
For impedance matching, just plug in the proper link. These new B & W plug-in links make your rig adaptable to practically any impedance as quickly as you can plug in a link with the correct number of turns. Type 3750 mounting bar for HDV coils, Type 3550 for TVH-TVL-BVL coils. Links available in 1-3-6 and 10 turns.

BARKER & WILLIAMSON, Inc.

QUIPMENT

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.

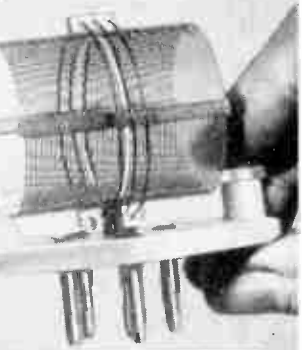


B & W BUTTERFLY VARIABLE CAPACITORS

Compactness and symmetry make these B & W JCX Variable Capacitors an outstanding favorite. Heavy rounded edge plates permit ratings to 2500 volts D.C. unmodulated and 1500 volts D.C. in modulated final amplifier circuits. They may be used as a capacitor-inductor assembly with any B & W "B" or "BX" inductors including the fixed or variable link types. Voltage rating measured at 30 megacycles.

B&W ALL-BAND FREQUENCY MULTIPLIER

Model 504—A fixed-tuned, 80- to 10-meter broadband 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. Ideal as a basic unit for your new rig.



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.

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QUALITY ELECTRONIC
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NEW!

CRYSTAL
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250 Kc. to 1000 Mc.

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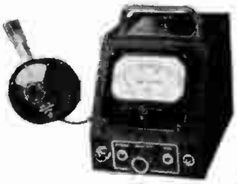
Model 112
300 Mc. to 1000 Mc.

STANDARD SIGNAL GENERATOR
Model 82
20 Cycles to 50 Mc.



MEGACYCLE
METER
Model 59
2.2 Mc. to 400 Mc.

STANDARD SIGNAL GENERATORS

MODEL	FREQUENCY RANGE	OUTPUT RANGE	MODULATION
65-B	75 Kc.-30 Mc.	0.1 microvolt to 2.2 volts	AM. 0 to 100% 400 cycles or 1000 cycle External mod., 50-10,000 c
78	15-25 Mc.; 195-225 Mc. 15-25 Mc.; 90-125 Mc. other ranges on order	1 to 100,000 microvolts	AM. 8200-400 cycle 625-400 cycle Fixed at approximately 3
78-FM	86 Mc.-108 Mc.	1 to 100,000 microvolts	Deviation 0-300 kc, 2 ra FM. 400-8200 cycles External modulation to 1
80	2 Mc.-400 Mc.	0.1 to 100,000 microvolts	AM. 0 to 30% 400 cycles or 1000 cycl External mod., 50-10,000
82	20 cycles to 200 Kc. 80 Kc. to 50 Mc.	0-50 volts 0.1 microvolt to 1 volt	Continuously variable 0- from 20 cycles to 20 K
84	300 Mc.-1000 Mc.	0.1 to 100,000 microvolts	AM. 0 to 30%, 400, 1000, o cycles. Internal pulse modu External mod., 50-30,000 c
90	20 Mc.-250 Mc.	0.3 microvolt to 0.1 volt	Continuously variable, 0 to Sinusoidal modulation 30 5 Mc. Composite TV modul

U.H.F. OSCILLATOR

MODEL	FREQUENCY RANGE	OUTPUT RANGE	OUTPUT IMPEDANCE
112	300 Mc. — 1000 Mc.	Maximum varies between 0.3 volt and 2 volts. Adjustable over 40 db range.	50 ohms

PULSE GENERATOR

MODEL	FREQUENCY RANGE	PULSE WIDTH	OUTPUT
79-B	60 to 100,000 cycles	Continuously variable from 0.5 to 40 microseconds	Approximately 150 volts po with respect to ground. " Output" 75 volts positive respect to ground.

SQUARE WAVE GENERATOR

MODEL	FREQUENCY RANGE	WAVE SHAPE	OUTPUT
71	Continuously variable 6 to 100,000 cycles	Rise time less than 0.2 microseconds with negligible overshoot	Step attenuator: 75, 50, 25 10, 5 peak volts fixed and 2.5 volts continuously vari

U.H.F. RADIO NOISE and FIELD STRENGTH METER

MODEL	FREQUENCY RANGE	INPUT VOLTAGE RANGE
58	15 Mc. to 150 Mc.	1 to 100,000 microvolts in antenna. 1 to 100 microvolt semi-logarithmic output meter, balanced resistance att ator with ratios of 10, 100 and 1000 ahead of all tu

VACUUM TUBE VOLTMETERS

MODEL	VOLTAGE RANGE	FREQUENCY RANGE	INPUT IMPEDANCE
62	0-1, 0-3, 0-30 and 0-100 volts AC or DC	30 cycles to over 150 Mc.	Approximately 7 mmfd
62-U.H.F.	0-1, 0-3, 0-30 and 0-100 volts AC or DC	100 Kc. to 500 Mc.	Approximately 2 mmfd
67	.0005 to 300 volts peak-to-peak	5 to 100,000 sine-wave cycles per second	1 megohm shunted by 30 m

MEGACYCLE METER

MODEL	FREQUENCY RANGE	FREQUENCY ACCURACY	MODULATION
59	2.2 Mc. — 400 Mc.	Within ±2%	CW or 120 cycles fixed at proximately 30%. Provision external modulation

CRYSTAL CALIBRATOR

MODEL	FREQUENCY RANGE	FREQUENCY ACCURACY	HARMONIC RANGE
111	250 Kc. — 1000 Mc.	0.001 %	.25 Mc. Oscillator: .25-450 M 1 Mc. Oscillator: 1-600 Mc. 10 Mc. Oscillator: 10-1000 M

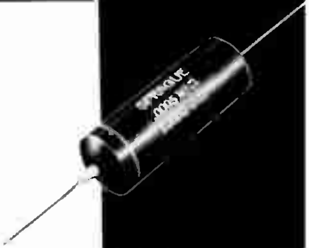
BRIDGES

MODEL	INDUCTANCE (L)	CAPACITANCE (C)	AC RESISTANCE (R)
102	0.5 microhenry to 110 henries	1 mmf. to 110 mfd. Power factor 0-30%	1 ohm to 11 megohms

MEASUREMENTS CORPORATION • BOONTON, N. J.

Just Right FOR YOUR RIG!

SPRAGUE CAPACITORS



HYPASS®—feed-through capacitors for bypassing harmonic currents in transmitters and for eliminating v-h-f interference from a-c mains and control circuits. Ideal for eliminating TVI ... developed to meet transmitter needs outlined by ARRL Headquarters.

FILTEROL®—line filters for suppressing "man-made" radio noises and television "scrambles" on practically any application. Small, self-contained, and easily installed, Filterols provide maximum noise suppression.

CERA-MITE★—the first complete line of miniature disc ceramics. Tough, dependable, and inexpensive. A type for every use—GA (General Application), high-K, and temperature-compensating.

METAL-ENCASED PAPER CAPACITORS—long a favorite with amateurs, these metal encased capacitors are impregnated and filled with KVO, the efficient dielectric material. Long-life and reliability are outstanding.

TELECAP★—phenolic molded paper tubulars assure maximum dependability under extreme heat, humidity, and shock. These Black Beauties are oil-impregnated in ratings from 600 to 12,500 volts.

ATOM®—metal-encased, dry electrolytics are engineered especially for tough television replacement applications. Small enough to fit anywhere, Atoms withstand punishing 85°C. (185°F.) temperatures.

TWIST-LOK★—dry electrolytics, sealed in aluminum cans with twist-prong tabs for washer or direct-to-chassis mounting. Top dependability in small size at 85°C. (185°F.). Sprague makes more ratings than any other manufacturer. Ideal for TV servicing.

MICAS—of maximum quality for R-F applications where high "Q" and high insulation resistance are required. There is a Sprague mica capacitor for every amateur requirement, from tiny half "postage stamp" types to giant ceramic-jacketed types.

KOOLOHM® RESISTORS

The only resistors wound with ceramic coated wire and doubly protected by a glazed ceramic housing and unique end seal to guard against moisture. Wires cannot short. Koolohms can be mounted anywhere, even flat against a chassis or grounded parts. They can be used safely at full wattage ratings.

WRITE FOR INTERESTING SPRAGUE LITERATURE COVERING THESE PRODUCTS OR SEE YOUR RADIO PARTS SUPPLIER.

★TRADEMARK

SPRAGUE®
PIONEERS IN ELECTRIC AND ELECTRONIC DEVELOPMENT

SPRAGUE PRODUCTS COMPANY
(Distributors' Division of Sprague Electric Co.)
North Adams, Massachusetts

WORKSHOP antennas

A Complete Line for all High-Frequency Bands

AMATEUR
TELEVISION
FM
EMERGENCY SERVICES

RADIO RELAY
MICROWAVE RELAY
STL RELAY
TV RELAY

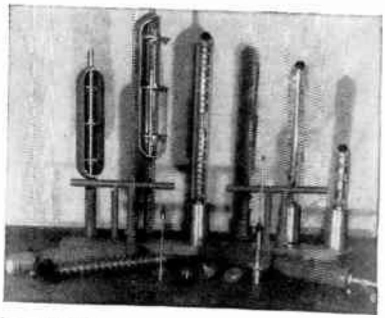
Government

RADAR
GUIDED MISSILE or ROCKET-BORNE
COMMUNICATION and NAVIGATION

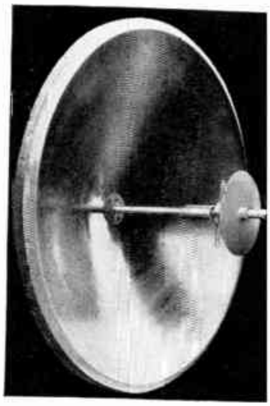
The WORKSHOP *pioneered* antennas for the high-frequency spectrum, and today our complete line covers all bands from 28 mc. up. Our standard antennas fit the majority of applications, but where special situations must be met, slight modifications in design usually accomplish the desired result with a minimum of time and expense.

Engineering and Contract Service. The WORKSHOP handles scores of special government and commercial antenna problems every year from design through production. With our new plant and greatly expanded facilities, we are able to handle a larger proportion of this type of contract work than ever before. If your product or service requires high-frequency antennas — research, design, test, or production — get in touch with the WORKSHOP. Write, or phone Needham 3-0005. No obligation.

THE WORKSHOP ASSOCIATES, Inc.
Specialists in High-Frequency Antennas
139 Crescent Road, Needham Heights 94, Mass.



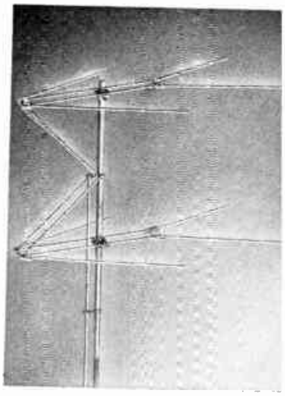
A representative group of h
precision radar antennas.



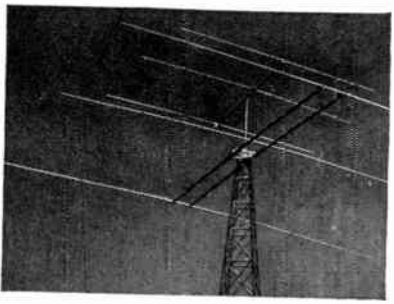
Standard Parabolic Relay Anten
for Studio-to-Transmitter Link
920-960 mc. band.

Left — High-Gain Beacon Anten
Recommended by all 152-162
equipment manufacturers. Hundre
are in use throughout the countr

Below — WORKSHOP DUBL-V
Television Antenna, double-stacke



Below — 10 over 20 Stacked A
ray — the last word in amateu
antenna equipment.



DO YOU KNOW THIS ABOUT ASTATIC MICROPHONES?

THE CRYSTAL ELEMENTS of Astatic's famous D-104 and the long-popular T-3 Microphones now have special METALSEAL protection against moisture and dryness.

THE NEWEST MODEL in the Astatic Microphone line is the Synabar, Model DR-10, a unidirectional cardioid crystal microphone of highest performance quality. An outstanding feature is the use of a special sintered metal to cancel out 15 db front to back, making the Synabar, for practical purposes, dead to sound from rear. Output level —54 db, frequency response 50 to 10,000 c.p.s. Has response selector switch, METALSEAL protection of crystal elements.

Model D-104

Model T-3

Model DR-10-S

WRITE FOR COMPLETE DETAILS

THE Astatic CORPORATION
CONNEAUT, OHIO
IN CANADA CANADIAN ASTATIC LTD. TORONTO ONTARIO

Astatic Crystal Devices manufactured under Brush Development Co. patents

VOLT-OHM-MIL-AMMETERS

for your every need

★ Note . . . These Sensational Improvements

TRIPLETT Model 630

- ★ Large 5½ Inch Meter in special molded case under panel.
- ★ Resistance Scale Markings from .2 Ohms to 100 Megohms. Zero Ohms control flush with panel.
- ★ Only one switch. Has extra large knob 2½" long, easy to turn, flush with panel surface.
- ★ New molded selector switch, contacts are fully enclosed.
- ★ All resistors are precision film or wire wound types, all sealed in molded compartment.
- ★ Unit construction—Resistors, Shunts, Rectifier, Batteries all are housed in a molded base built right over the switch. Provides direct connections without cabling. No chance for shorts.
- ★ Batteries easily replaced—New Double Suspended Contacts.
- ★ Ranges: DC Volts 0 to 6000, 20,000 Ohm/Volt; AC Volts 0 to 6000, 5000 Ohm/Volt; DB: − 30 to +70; Direct Current from 0–60 Microamp. to 0–12 Amps; Resistance: 0–1000–10,000 Ohms, 0–1–100 Megs.

ABSORPTION FREQUENCY METER

Model 3256

A band-switching, tuned absorption type frequency meter that covers five amateur bands. Has Germanium 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. 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. Coil is removable and other coils may be substituted for special bands. Useful for checking: Fundamental frequency of oscillating circuits; Presence, order and relative amplitude of harmonics, Parasitic oscillations, etc. Size: 7½" x 2½" x 2¼". Metal case with gray enamel finish, black trim.

POCKET SIZE Model 666-R

For your AC and DC Voltage, Direct Current and Resistance analyses to 3 Megohms. Enclosed selector switch of molded construction keeps dirt out. Retains contact alignment permanently. Unit Construction — All Resistors, shunts, rectifier and batteries housed in a molded base integral with the switch. Eliminates chance for shorts. Direct connections. No cabling. All precision film or wire-wound resistors, mounted in their own compartment—assures greater accuracy. 3" RED•DOT Lifetime Guaranteed instrument. Red and black markings on a white background. Easy to read scale. Precalibrated rectifier unit. Self-contained batteries.

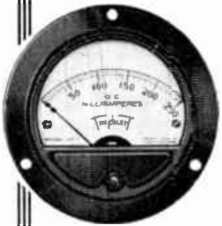
RANGES: AC-DC Volts: 0–10–50–250–1000–5000, 1000 Ohms/Volt; Direct Current: 0–10–100 Ma., 0–1 Amp.; Resistance: 0–3000–300,000 Ohms, 3 Meg. Black molded case, completely insulated, 3¹⁄₁₆" x 5⅞" x 2 ⁹⁄₁₆". White panel markings.

Model 666-HH VOLT-OHM-MILLIAMMETER

A complete miniature laboratory for DC-AC Voltage, Direct Current and Resistance analyses. The answer to V-O-Ma. requirements of radio servicemen, amateurs, industrial engineers, etc. Greater scale readability on the 3" RED•DOT Lifetime guaranteed instrument with red and black scale markings. Simplified switching.

RANGES: AC-DC Volts: 0–10–50–250–1000–5000, 1000 Ohm/Volt; DC Ma: 0–10–100–500; OHMS: 0–2000–400,000. Self-contained plug-in batteries. Black molded case, completely insulated, 3¹⁄₁₆" x 5⅞" x 2⁹⁄₁₆". Panel with white markings.

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, thermo-ammeters, DB meters, VU meters and electrodynamometer type instruments.

Address all inquiries to Dept. Q–51

In Canada:

Triplett Instruments of Canada, Georgetown, Ont.

FOR THE MAN WHO TAKES PRIDE IN HIS WORK

Triplett

TRIPLETT ELECTRICAL INSTRUMENT COMPANY • BLUFFTON, OHIO, U.S.A.

Collins for top performance

Collins 75A-2 Receiver

The new Collins 75A-2 is a double-conversion superheterodyne, designed specifically to give top performance for the amateur in the 160, 80, 40, 20, 15, 11 and 10 meter bands. It features sensational stability, calibration accuracy and sensitivity, and provides the greatly improved selectivity required for successful ham operation under present overcrowded conditions.

This increased selectivity is provided by nine tuned circuits at 455 kc i-f, plus an improved crystal filter which is variable by a front panel control. When the 75A-2 is tuned, the increased skirt selectivity is instantly apparent. There are interference-free holes in the crowded phone bands. High-pitched heterodynes are practically eliminated. This leaves the phasing control free for use in eliminating low-pitched heterodynes. The range of the phasing control notch has been extended downward to approximately 200 cps.

Only the band in use is visible. The slide rule dial is calibrated directly in one-tenth mc, the vernier dial at one kc intervals on all bands except 11 and 10, where it is two kc. A vernier zero set control is on the front panel. Other front panel controls include band change; on-off-standby; CW—AM—FM; r-f gain; audio gain; limiter in-out and calibrate; separate CW limiter; selectivity; crystal filter phasing; BFO pitch; antenna trimmer.

The 70E-12 PTO employs a new Collins permeability tuned two-tube circuit, which assures improved stability unaffected by variations in tubes.

15 miniature tubes and rectifier are employed: 6AK5 r-f amplifier, 6BE6 h-f mixer, 12AT7 crystal oscillator, 6BE6 l-f mixer, two 6BA6 VFO's, three 6BA6 i-f amplifiers, 6AL5 AVC—detector—audio detector, 6BA6 BFO, 6AL-5 noise limiter, 12AX7 AVC amplifier—audio amplifier, 6AL5 CW noise limiter, 6AQ5 power amplifier, and 5Y3 power rectifier.

The 8R-1 100 kc crystal calibrator and the 148C-1 NBFM adapter, shown on this page, are available as accessories, for plugging into completely wired sockets on the top of the chassis. The operation of both units may be controlled by switches located on the front panel.

8R-1

148C-1

75A-2 dimensions:
21 1/8" wide, 12 7/16" high, 13 5/16" deep.
Power source: 115 volts 50/60 cycles a-c.
Shipping weight: 70 lbs.
Net domestic prices:
75A-2 receiver: $420.00
10-inch speaker in matching cabinet: $20.00
8R-1 crystal calibrator: $25.00
148C-1 NBFM adapter: $22.50

For the best in radio communications, its . . . COLLINS

COLLINS RADIO COMPANY, Cedar Rapids, Iowa

11 W. 42nd Street, NEW YORK 18
2700 W. Olive Avenue, BURBANK

Collins for more QSO'S

Collins
KW-1 Transmitter

This new kilowatt rig is as convenient to operate as a Collins 32V. Completely integrated in an attractive wrinkle-finish cabinet, it is a product of the most modern design techniques. The power input is 1000 watts on CW, phone, and NBFM. Provision for NBFM is built in as standard equipment. The frequency range covers the 160, 80, 40, 20, 15, 11 and 10 meter amateur bands.

Complete bandswitching of the exciter, driver, and power amplifier is accomplished by a single control on the front panel. This reduces to four the number of tuning functions required: bandswitch selection, frequency setting, PA tuning, and PA loading. Over any narrow frequency range, it is only necessary to adjust the frequency control. Frequency control is by means of a newly developed, extremely stable, hermetically sealed master oscillator. Operating controls can all be reached from a sitting position if the KW-1 is installed next to the desk. However, the entire exciter-power amplifier section may be removed and placed on the desk if desired.

TVI reduction is accomplished by the use of multiple-tuned circuits at the output frequency on every band. A minimum of three circuits at the output frequency greatly attenuates not only the second and third harmonics, but also sub-harmonics. Particular attention has been paid to filtering all control and power leads entering the exciter-power amplifier compartment, which itself is a totally enclosed and shielded structure. A 35C-1 low pass filter is incorporated as standard equipment. The output network is a conventional pi followed by an L section for increased harmonic attenuation.

The speech amplifier has a peak clipper plus a low and high level filter, and permits high-percentage modulation without splatter.

Tube line-up: Exciter — one 6BA6, five 6AQ5's, one 807W, one 12AU7, one 6AL5. Power amplifier — two 4-250A's. Speech amplifier — one 12AX7, one 6AL5, two 12AU7's, two 6B4G's, two 810's. Rectifiers — two 872's, one 5R4GY, three 5V4's.

Metering includes modulator current, PA plate current, high voltage, line voltage, multi-purpose meter, and antenna ammeter. Line fuses, plus overload relay in Class C amplifier current lead, provide circuit protection.

Dimensions:
28" wide, 18" deep, 66 1/2" high.

Power source:
115/230 volts, 60 cycle single phase grounded neutral.

Price to be announced.

Collins 32V-2 Transmitter

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 heart of the 32V-2 is the Collins 70E-8A permeability tuned oscillator, used as the VFO. The calibration is very accurate, and stability compares favorably with most crystals used by amateurs.

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 OA2 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:
21 1/8" wide, 12 7/16" high, 13 7/8" deep.
Power source: 115 volts 50/60 cycles a-c.
Shipping weight: 128 pounds.
Net domestic price **$575.00**

35C-1 LOW PASS FILTER

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. 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.

Net domestic price **$40.00**

148B-1 NBFM ADAPTER

The Collins 148B-1 Narrow Band FM Adapter is for use with either the 32V-1 or the 32V-2 amateur transmitter. 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.

Net domestic price **$31.75**

Collins for all communications

20V one KW
AM broadcast transmitter

The advanced engineering, reliability, and high performance typical of Collins ham gear are also characteristic of Collins airborne and ground-based radio communication and navigation equipment, and Collins AM and FM broadcast station transmitters, audio equipment and accessories.

Collins fixed based transmitters are available in powers ranging from 75 watts to 50 kilowatts, CW, FSK and phone, operating at high and very high frequencies, autotuned, servo tuned and manual tuned. Ground station receiving gear includes single and multichannel and continuously tunable rack mounting equipment for A1, A2, and A3 reception at high and very high frequencies.

Collins airborne transmitters and receivers provide facilities for HF and VHF two-way phone communication, for omnirange navigation, and for instrument landing.

For broadcast stations, Collins makes a wide range of transmitters, as well as a complete line of speech equipment.

All Collins equipments for commercial and personal applications are designed and made to the exacting standards set for military applications. Our engineering and manufacturing personnel do not know how to shortcut or cheapen.

Collins 51J Communication Receiver

Left to right, Collins 51R-2 VHF navigation receiver, control box, omni-bearing selector, deviation indicator, and radio magnetic indicator.

Collins 51V-1 UHF Glide Slope Receiver

Collins 37J-1 VHF Aircraft Antenna

Collins 37P-1 UHF Glide Slope Antenna

Now! AMPHENOL AUTO-DIAL ROTATOR

THE PRACTICAL ROTATOR FOR 2 AND 6 METERS

'Auto-Dial" is the Rotator for VHF! Conserva-ive in price and size yet rugged enough for mateur use.

Hams will appreciate the fast, automatic ction. No buttons or switches to hold while rray rotates. Just turn knob to desired direc-ion and continue QSO. Antenna rotates in one irection to exact number of degrees and tops! No coasting, no backlash!

UNUSUAL FEATURES OF "AUTO-DIAL"

- Slip-ring contacts of coin-silver—no line unbalance, no twisted feed-lines!
- Antenna rotates rapidly—only 22 seconds for complete revolution!
- Rotation in steps of 6 degrees permits exact orientation—accurate antenna field strength measurements!
- Lifetime lubricated! Sealed against dirt and moisture!
- Heavy Duty Motor!
- Inline mast mounting. Takes mast sizes 1" to 1½" O.D.

Rotator housing is cold-rolled steel, copper flashed for durability. Baked-on enamel finish. Weighs just 6 lbs.

Ingenious control unit shows at a glance direction the array is beamed.

AMPHENOL RF CONNECTORS

Unsurpassed for mechanical design and electrical efficiency. Provide lowest loss continuity in critical RF circuits with little or no impedance change or increase in standing wave ratio.

Wide variety of AMPHENOL RF Connectors available includes Plugs, Jacks, Receptacles, Adapters, etc. High and low voltage types, various impedances, many weatherproof or pressurized. All AMPHENOL connectors meet rigid government specifications.

AMPHENOL COAX AND TWINAX

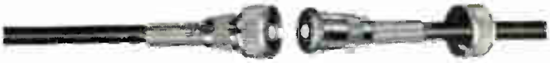
Produced to standards surpassing military specifications for electrical performance, mechanical excellence.

Use of AMPHENOL coax is a great step toward elimination of TVI!

Most AMPHENOL RG cables have polyethylene dielectric for low loss, flexibility, mechanical stability. Certain AMPHENOL cables utilize TEFLON and withstand temperatures as high as 500° F.

AMPHENOL can supply coax and Twinax in a large number of types.

To make connection, unscrew coupling ring from one cable end, slip it back over cable.



You now have male and female cable ends. Screw the remaining ring on to the ringless end...

...tighten coupling ring. Contact is made, the junction is completed without loss of time or effort.

AMPHENOL MICROPHONE CONNECTORS

AMPHENOL manufactures an extensive line of connectors to fit practically all makes of microphones.

The 75-MC1F Microphone Connectors, illustrated above, function as either male or female fitting so that in use a mating connection is always ready for instant application.

Distinctively styled, AMPHENOL'S 75-MC1F, single contact, shielded cable type microphone connectors are made of chrome plated, machined brass. Will accommodate cables up to ¼" diameter.

The 75 Series connectors include Jacks, Plugs, Receptacles, "Mike" Switches, etc. See them at your dealer.

The 80 Series, single and double contact connectors are designed for shielded cables and have many uses in both audio and RF circuits. Obtainable as male or female cable connectors or chassis units.

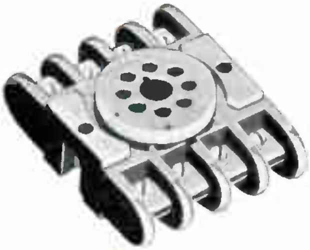
The 91 Series includes both three and four contact connectors, polarized to prevent incorrect insertion. Procurable as plugs, cable jacks and chassis receptacles, either male or female types.

AMPHENOL

AMERICAN PHENOLIC CORPORATION • 1830 SO. 54th AVE., CHICAGO 50, ILLINOIS

AMPHENOL

INDUSTRIAL TUBE SOCKETS



Peak performance — utmost dependability! These sockets have rugged insulating barriers, removable contacts, RMA numbered reversible screw type terminals to simplify wiring and permit use of wiring harness and terminal lug connections. Illustrated is 146-103 of molded Melamine. Other models such as barrier type high voltage sockets, barrier type miniature 7-pin sockets, stair type sockets for jumbo tubes, high voltage mounting plate type sockets and stilt type tube sockets are also procurable in Melamine, Steatite and high grade phenolics.

PATENTED "CLOVER LEAF" CONTACTS ON ALL AMPHENOL INDUSTRIAL SOCKETS

The "Clover Leaf" contact provides four full lines of contact along each tube pin, assuring a high degree of rententivity with a contact resistance considerably less than .002 ohms. Contacts plated to resist corrosion.

AMPHENOL

MINIATURE 7 & 9 PIN SOCKETS

Available for every application. Materials used are finest available, include black bakelite, mica-filled bakelite, AMPHENOL Steatite and AMPHENOL'S own Ethylon-A with its high "Q" factor and low-loss properties. Also Zip-In sockets for high speed production.

AMPHENOL MIP

Unequalled in strength, versatility and appearance. Steel mounting plate molded into bakelite body. Available in black bakelite or mica-filled bakelite in wide variety of contact arrangements. Compact MIP sockets also obtainable for 8 pin Octal and Loktal tubes.

AMPHENOL "S" SOCKETS AND "CP" PLUGS

Combine the convenience of AMPHENOL Retainer Ring design with the inherent high quality of AMPHENOL Steatite.

Mount without screws or rivets, withstand extremely high temperatures. Silver-plated phosphor bronze contacts. Also available in black bakelite or mica-filled bakelite.

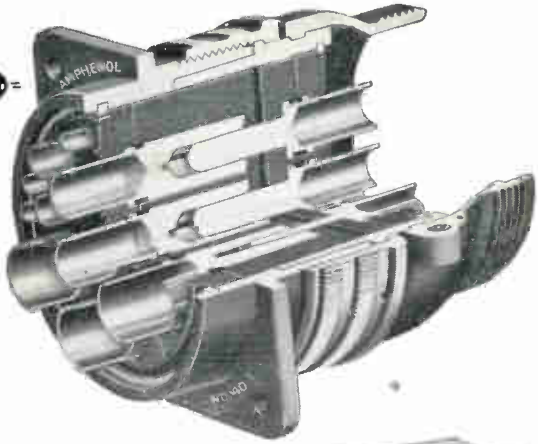
AMPHENOL STEATITE

AMPHENOL manufactures a comprehensive line of tube sockets made of AMPHENOL Steatite. Suited for applications where low-loss and highest degree of efficiency are required.

Positive moisture resistance is assured through vacuum impregnation with Dow-Corning #200.

COMPLETE CABLE HARNESSES AND ASSEMBLIES AVAILABLE!

AMPHENOL produces complex wired assemblies and harnesses involving many components as one quality unit—reducing procurement, production planning, inventory control and component inspection costs.



AMPHENOL "AN" CONNECTORS

For Power, Signal and Control Circuits in Aircraft and Electronic Equipment

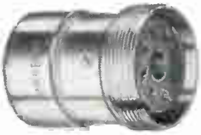
AMPHENOL leads the industry in the manufacture of "AN" connectors to meet government specifications under MIL-C-5015.

Available in many shell styles, AMPHENOL "AN" connectors feature many improvements worked out by AMPHENOL engineers in cooperation with government engineers.

Included among these superior features are:

- Lowest milivolt drop.
- Coupling rings machined from solid aluminum bar stock. Extra high tensile strength (53,000 pounds).
- AMPHENOL non-rotating contacts for easy, fast soldering.
- Coupling rings and assembly screws drilled for safety wiring.
- Simple assembly, no special tools required.

In AMPHENOL'S Catalog 74 will be found the most complete listing of "AN" connectors available from a single manufacturer under Specification MIL-C-5015. All inserts and shell types available from regular AMPHENOL production.

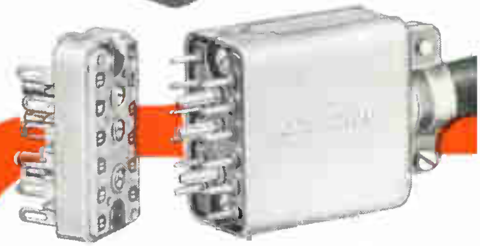


HEAVY DUTY RADIO CONNECTORS

Compact, lightweight, used extensively for connecting various units of transmitters and testing apparatus and as power connectors for mobile transmitters and receivers. Completely encased in heavy drawn brass cadmium plated shell. Entirely free of shock hazard—will not radiate RF. Polarized shell permits 4 different element positions for added circuit protection. Plugs, jacks and receptacles available in 4, 5, 6, 8 and 12 contacts.

"See Catalog 74 for complete description and listing of these and other AMPHENOL products."

AMPHENOL



RACK AND PANEL CONNECTORS
26 SERIES

Connectors obtainable with 11, 15 and 20 contacts. All have eyelets inserted in mounting holes for added strength, holes for wiring, and interlocking barriers to prevent accidental shorting. Can be supplied with or without protective can and cable clamp as shown. Voltage rating 500 volts, 60 CPS at sea level. Mounting screw spacing on 11 contact, .864; on 15 contact, 1.188; on 20 contact, 1.620.

RELAY PLUGS
157 SERIES

Until the 157 Series was perfected, there were no rubber sealed connectors for sealed relays which could meet rigid MIL-C-5015 specifications. Tests show the 157 Series exceeds requirements, have NO measurable leakage rate during and after temperature cycling. Inserts employing the new AMPHENOL 1-501 thermosetting plastic dielectric for the front and rear with high quality resilient dielectric sandwiched between to provide required seal. Pressure seal maintained indefinitely. 157 Series available in standard "AN" insert arrangements, mate with conventional AN-3102 receptacles. Obtainable in Hex nut and solder types.

AMERICAN PHENOLIC CORPORATION

1830 SO. 54th AVE., CHICAGO 50, ILLINOIS

WHEREVER THE CIRCUIT SAYS —/\/\/\—

ADVANCED TYPE BT RESISTORS

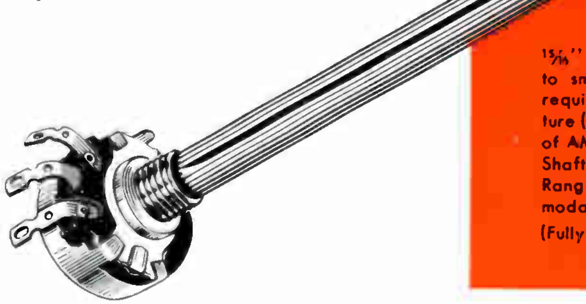
New type BT Insulated Composition Resistors—meet JAN-R-11 Specifications at ⅓, ½, 1 and 2 watts. Small size BTB specially designed for miniature 2 watt requirements. Type BT's are suited to television and similar exacting circuits. Extremely low operating temperature. Excellent power dissipation. 330 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

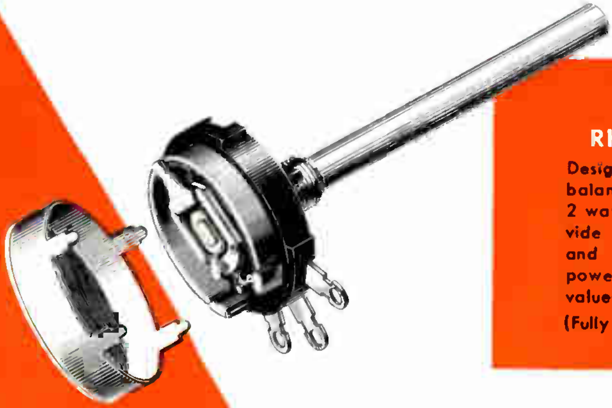
(Fully described in Catalog RDC8.)



NEWLY DEVELOPED TYPE Q VOLUME CONTROL

15/16" 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. Knab 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 9/16". Resistance values: 2 ohms to 10,000 ohms.

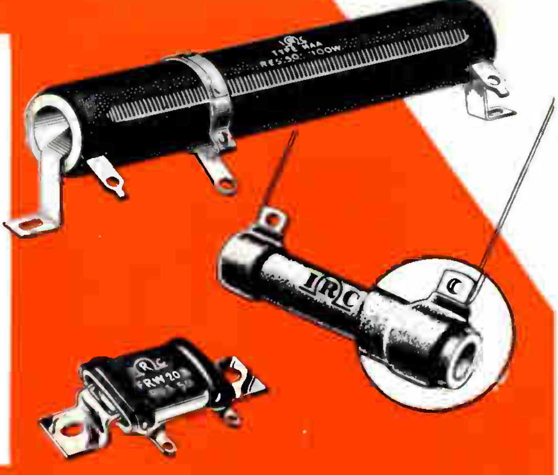
(Fully described in Catalog RDC1-A.)

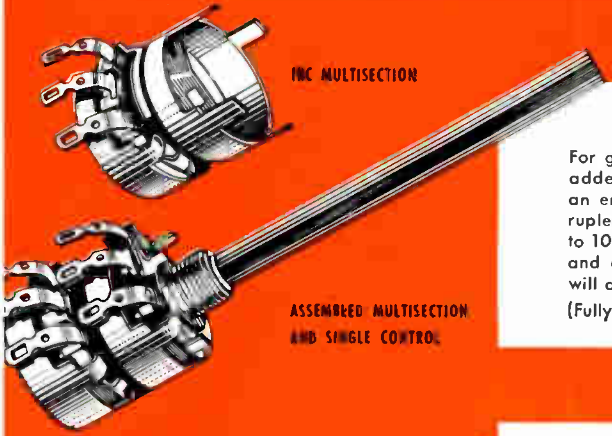
POWER WIRE WOUND RESISTORS

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.)



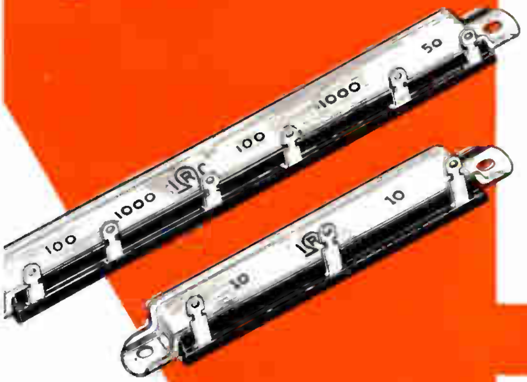

IRC MULTISECTION

ASSEMBLED MULTISECTION AND SINGLE CONTROL

Multisections

For ganged controls, IRC MULTISECTIONS are added to Q controls like switches to provide an endless variety of duals, triples and quadruples. Available in 17 values from 1000 ohms to 10 megohms. MULTISECTIONS are as easily and quickly attached as switches—and duals will accommodate Type 76 switches.

(Fully described in Catalog RDC1-A)



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

WHEREVER THE CIRCUIT SAYS -/\/\/-

CLOSE TOLERANCE DEPOSITED CARBON PRECISTORS

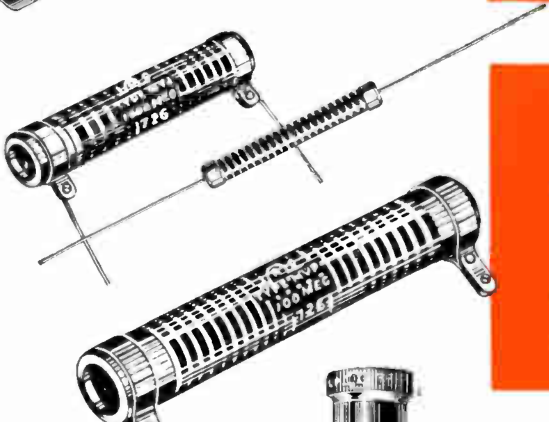
PRECISTORS offer a unique combination of close tolerance, stability and economy. Pure crystalline carbon bonded to selected ceramic cores overcomes limitations of carbon composition resistors and higher cost of precision wire wounds. PRECISTORS offer wide range of values, guaranteed accuracy, high stability, low voltage coefficient, excellent frequency characteristics, predictable temperature coefficient.

(Fully described in Catolog RDC-3.)

HIGH FREQUENCY RESISTORS

Type MP Resistors ore designed for frequencies above those of conventional resistors. 2 wotts to 90 watts. Special construction, with resistance film bonded to steatite ceramic form, provides stable resistors of low inductance and capacity. Type MPM's ore miniature ¼ wott units for small-space, high frequency receiver applications.

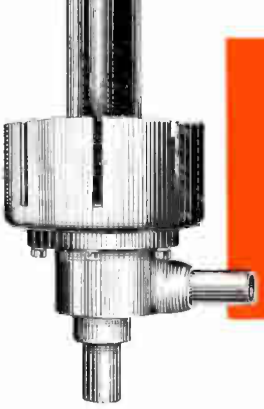
(Fully described in Cotolog RF-1.)



HIGH VOLTAGE RESISTORS

Type MV's meet high resistance and power requirements in high voltage applications. Resistance coating in helical turns on ceromic 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.

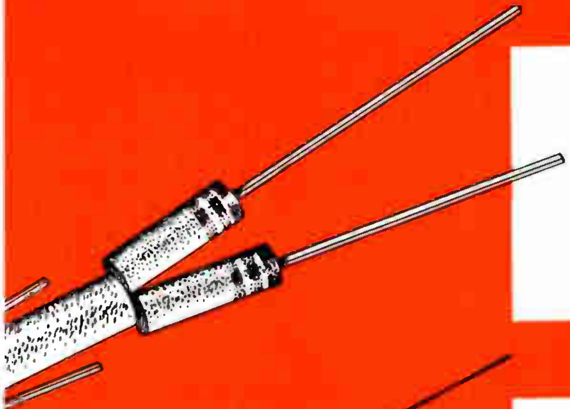
(Fully described in Catalog RF-2.)

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.

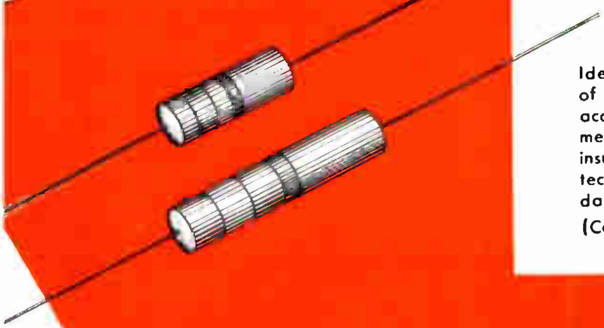
(Fully described in Catalog RD-2.)



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

401 N. Broad Street Philadelphia 8, Pa.

In Canada: International Resistance Co., Ltd., Toronto, Licensee

RCA tube line-ups for 1951 . . .

CW, FM, and Phone—160, 80, 40, 20, and 10 Meters

Input Watts to Final							
CW & FM	PHONE	OSC.	MULTIPLIERS	BUFFER	FINAL	MODULATORS	CLASS
17	15	6AK6	6AK6	6AK6	5763	6AQ5 6AQ5	AB1
40	27	6AG7	6AG7	6AG7	2E26	6L6 6L6	AB1
75	60	6AG7	6AG7	6AG7	807	6L6 6L6	AB1
120	90	6AG7	6AG7	5763	829-B	807 807	AB2
260	175	6AG7	6AG7	2E26	812-A	807 807	B**
345	260	6AG7	6AG7	6AG7	4-65A	811-A 811-A	B
500	375	6AG7	6AG7	2E26	4-125A/ 4D21	811-A 811-A	B
500	400	6AG7	6AG7	2E26	813	811-A 811-A	B
520	350	6AG7	6AG7	807	2-812A	811-A 811-A	B
750	500	6AG7	5763	807	810	811-A 811-A	B
1000	1000	6AG7	807	812-A	833-A	810 810	B
1000	800	6AG7	6AG7	807	2-813	810 810	B

CW, FM, and Phone—VHF Bands

Input Watts to Final							
CW & FM	PHONE	BANDS UP TO	OSC.	MULTIPLIERS/BUFFER	FINAL	MODULATORS	CLASS
18*	12*	1700 Mc	6AK6	5763 5876	2-5876	6AQ5 6AQ5	A
17	15	175 Mc	6AK6	6AK6 5763	5763	6AQ5 6AQ5	AB1
40	27	160 Mc	6AK6	6AK6 5763	2E26	6L6 6L6	AB1
50	36	200 Mc	6AK6	6AK6 5763	832-A	6L6 6L6	AB1
150	120	200 Mc	6AK6	5763 2E26	829-B	807 807	B**

*Watts at 500 Mc

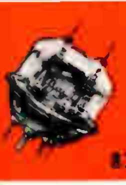
**Special triode connection

selected to save you time and calculation


810


811-A


812-A


832-A


833-A


2E26


807


813

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
833-A	10.0	10	35	30		1500
			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
829-B	6.3 12.6 (H)	2.25 1.125	9#	200	7*	120*
832-A	6.3 12.6 (H)	1.6 0.8	6.5#	200	5*	50*
5763	6.0(H)	0.75	16#	175	2	17
			RCA TETRODES			
4-65A	6.0	3.5	5#	50	10	350
4-125A/ 4D21	5.0	6.5	6.2#	120	20	500
			RCA RECTIFIERS			
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.			

ICAS—Intermittent Commercial and Amateur Service
#Control-grid to screen-grid mu-factor
*Total for tube


829-B


5763


5876


4-65A


4-125A/4D21


866-A


816


5R4GY

RCA has a popular tube for every Amateur service, every power, and every active band.

In addition, there are special-application types, such as *phototubes, acorns, kinescopes, iconoscopes,* and the well-known *receiving types* in metal, glass, miniature, and subminiatures.

For additional technical data on these RCA tube types, see your local RCA Tube Supplier, or write RCA, Commercial Engineering, Section A35M, Harrison, N. J.

Are you getting the new RCA HAM TIPS? There's a copy waiting for you at your RCA Tube Supplier.

RADIO CORPORATION of AMERICA

ELECTRON TUBES

HARRISON, N. J.

Oui-Si-Ja-Da-Ding Hao

Yes, say hams the world over it's

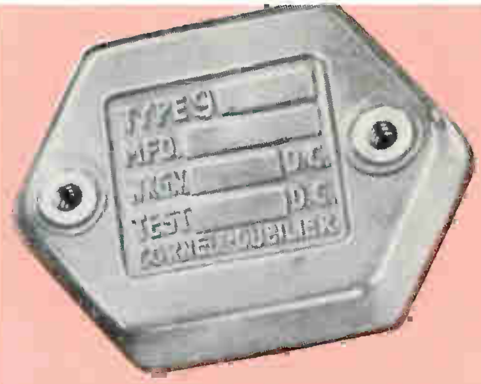
CORNELL DUBILIER

CAPACITORS • VIBRATORS • ROTATORS • ANTENNAS • CONVERTERS

for dependable service in all radio and TV applications

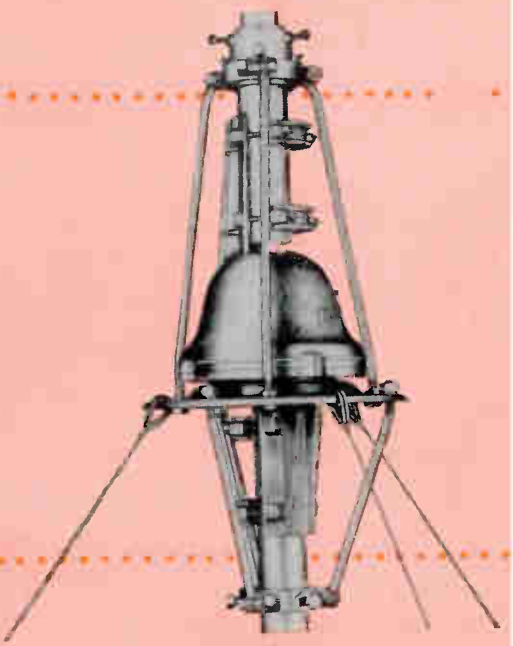
Whenever hams CQ, Cornell-Dubilier gets the recommendations on dependable capacitors; rotator, TV, FM and AM antennas; power converters and vibrators. For the most for your money look for the C-D quality trade-mark on these components. Cornell-Dubilier Electric Corporation, Dept. AH50, South Plainfield, N. J. Other plants in New Bedford, Brookline and Worcester, Mass.; Providence, R. I.; Indianapolis. Ind.; and subsidiary, The Radiart Corp., Cleveland, Ohio.

1910 CORNELL-DUBILIER PRODUCTS CD 1950



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" ROTATORS

Rotators that will handle 10 meter and 20 meter beam antennas. Ruggedly constructed, these weatherproof rotators will fit masts up to 2 inches O.D. Furnished with most accurate compass indicator control device available. Also FM, AM and TV antennas of all types, shapes and sizes.

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.

MODEL
T-602

Here's the exact duplicate of the TEC Projection Oscilloscope developed for the U. S. Navy for mass electronics training. Makes waveforms brilliantly clear to groups as large as 750 persons! No more students hunching round a tiny image! No more mistaking what you mean! Only TEC gives you such advanced features for top performance and flexibility:

IDEAL FOR

- SCHOOLS AND COLLEGES
- MILITARY TRAINING
- LABORATORIES
- INDUSTRY TRAINING
- DEMONSTRATIONS
- GOVERNMENT AGENCIES

TEC *Electronic* BLACKBOARD

AMPLIFIER CHARACTERISTICS

PARAMETER	Y-AXIS	X-AXIS	Z-AXIS
A-C Gain	1 mv rms/inch	60 mv rms/inch	2 volts rms visible
D-C Gain	2.5 volts/inch	—	—
L-F Response	± 10% 2 cps	± 10% 2 cps	± 10% 2 cps
H-F Response	± 10% 750 kc	± 10% 325 kc	± 10%
	— 3 db 825 kc	— 3 db 700 kc	100 kc
Input Impedance	2 megohms, 30 mmf	2 meg 30 mmf	1 megohm, 20 mmf
Attenuator	1, 10, 100 X	1, 10, 100 X	—
Defl. Plates—Direct	A-C, 1.5 megs, 20 mmf	A-C, 1.5 megs, 20 mmf	—

SWEEP CIRCUITS

Recurrent sweeps:	1 cps to 50 kc, auto. retrace blanking circuit
Driven sweeps:	20 us to 10^6 us, auto. brightening circuit
Linearity:	± 10% variation of spot velocity
Synchronization:	Internal, external, power line; ± input, high gain
Line freq sweep:	60 cps nominal, with phase shifter

GENERAL

Size: 26" wide, 33" deep, 66" high; weight: 350 lbs.
Input: 105 - 130 volts, 50/60 cps, 600 watts
Price: $2495 - F.O.B. New York City

PICTURE SIZE

Integral screen:
18" x 24", non-directional

Projected screen:
8' x 10', or larger, as desired

SIGNAL CALIBRATOR

Built-in, line-frequency signal, range 0 - 10 volts. Reads peak-to-peak voltage

OPTICS

5RPA tube, 20 kv acceleration, f/1.9 - 5" Bausch & Lomb coated lens. Tube brightness (100 x 100 line raster) 130 foot candles, average

NOTE: Model T-602 available as a medium gain wide-band instrument, on special order. Price on application

TEC TELEVISION EQUI

238 WILLIAM STREET • N. Y. C. 7, N. Y. • CORTLANDT 7-5160

TEC

MODEL T-601-A WIDE-BAND HIGH-GAIN OSCILLOSCOPE

The TEC Model T-601-A Oscilloscope is a general-purpose high-quality cathode-ray oscilloscope, which has been designed for laboratory engineering use. It has all the features requisite to a precision laboratory equipment, and is in every sense a professional unit. It uses a Type 5UP 5" C-R-T with an accelerating potential of 1750 volts. It includes the necessary engineering features of phaseable 60 cycle sweep, internal voltage calibrator, as well as wide band amplifiers and driven sweep circuit.

AMPLIFIER CHARACTERISTICS

PARAMETER	Y-AXIS
Gain	10 mv rms/in
L-F Response	10% tilt 10 cps
H-F Response	— 3 db @ 12 Mc
Input Z	2 Megs, 20 mmf
Attenuator range	1, 10, 100, 1000 X

X-AXIS	Z AXIS
250 mv rms/in	1.5 v rms visible
10% tilt 20 cps	—
— 3 db @ 250 kc	Useable to 1 Mc
2 Megs, 10 mmf	50 K ohms, 50 mmf

SIGNAL CALIBRATOR

Built-in, line frequency signal, range 0 - 10 volts.

Reads peak-to-peak voltage.

GENERAL

Range of Centering: 3" x 4" approximately X and Y axes respectively

Line Width: Approximately 0.5 mm at best focus

Input: 105 - 125 x rms 50/60 cycles; 175 watts

Size: 14" High x 10" Wide x 17" Deep

Weight: Approximately 45 lbs.

- 17 TUBES INCLUDING 5" CRT.
- 10 MILLIVOLT SENSITIVITY
- 12 MEGACYCLE BANDWIDTH
- DEFLECTION PLATES AVAILABLE ON TERMINAL BOARD
- CONTINUOUSLY VARIABLE CALIBRATOR
- SWEEP MAGNIFICATION 5 TIMES SCREEN SIZE
- GOOD TRANSIENT RESPONSE
- TRIED AND PROVEN CIRCUITS

SWEEP CIRCUITS

Recurrent Sweeps: 10 cps to 100 kc, automatic retrace blanking circuit

Driven Sweeps: 10 microsec. to 10^5 microsec., automatic brightening circuit

Linearity: ± 10% variation of spot velocity

Synchronization: Internal, external, power line ± inputs high gain

Sensitivity Minimum sync voltage 30 mv rms

To synchronize at 1000 cps

Sweep Expansion: 5 times full scale, undistorted

Maximum Writing Speed: 2.5 inches per microsecond

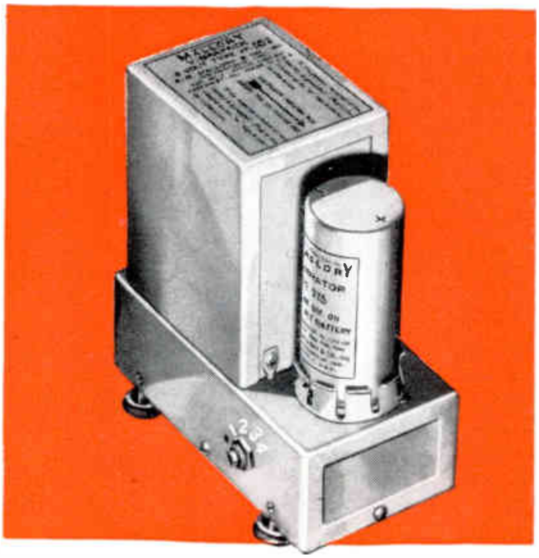
MENT CORPORATION

IN CANADA: THE AHEARN & SOPER CO. LTD., OTTOWA

You Can Rely on

P.R. MALLORY & CO. Inc. MALLORY

Approved Precision Products



VIBRAPACK* Power Supplies . . . A completely dependable source of high voltage where commercial AC is not available. A complete line includes units adaptable to many applications. High efficiency, low battery drain, rugged dependability, compact, light weight.

"HAMBAND" SWITCHES . . . This Ceramic Section Switch is one of a group of Mallory switches designed to meet your every need. Available in one or more sections with one or more circuits per section . . . ideal for band switching in low power transmitter circuits and many other applications.

You can rely on Mallory Approved Precision Products to give you long life and dependable service . . . fewer shut-downs and less trouble-shooting. You can rely on Mallory for helpful literature and advice on your technical problems. And you can rely, too, on your Mallory Distributor. He is ready to serve you.

Insist on Mallory Many amateurs have standardized on the complete line of Mallory Approved Precision Products . . . Capacitors . . . Controls . . . Inductuners . . . Resistors . . . Switches . . . Jacks & Plugs . . . Vibrators . . . Rectifiers . . . Battery Chargers . . . Noise Filters . . . *and a wide range of special items, accessories, tools and hardware.*

The Mallory Inductuner**

An excellent general purpose tuner for the entire frequency range from 52 through 216 megacycles. Write for technical information bulletin.

P. R. MALLORY & CO., Inc.
INDIANAPOLIS 6, INDIANA

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*Reg. U. S. Pat. Off.
**Reg. trademark of P. R. Mallory & Co., Inc. for inductance tuning devices covered by Mallory-Ware patents.

"THE HAM TUBE YOU WANT—

when you want it—

you'll find it in G.E.'s *complete* line!

. . . plus circuit-application help that I'm glad to give you regularly in Ham News, or by letter upon request."

Lighthouse Larry

© 1951 by General Electric Company

G-E POWER TUBES

GL-2C40
GL-2C43
GL-2E24
GL-2E26
GL-4D21/4-125A
GL-4-250A/5D22
GL-35T
GL-100TH
GL-203-A
GL-211
GL-592
GL-802
GL-803
GL-805
GL-806
GL-807
GL-810
GL-811-A
GL-812-A
GL-813
GL-814
GL-815
GL-826
GL-828
GL-829-B
GL-832-A
GL-837
GL-838
GL-1613
GL-1614
GL-1619
GL-1623
GL-1624
GL-1625
GL-8000
GL-8005
GL-8012-A
GL-8025-A

G-E RECEIVING TUBES

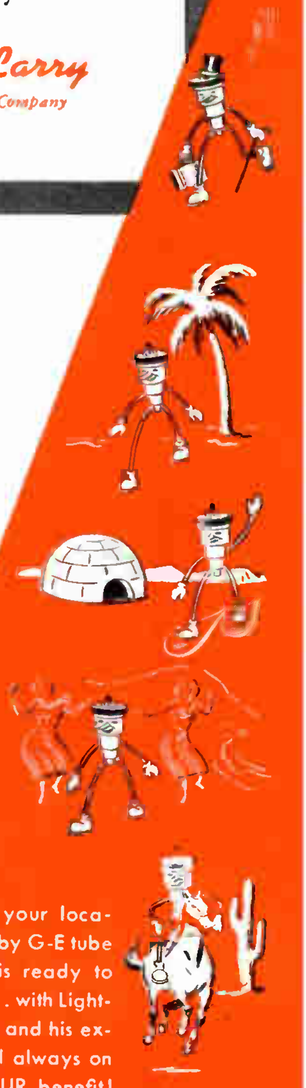
More than 500 types, covering every circuit need. Metal, glass, miniatures.

G-E RECTIFIER TUBES

5R4-GY
GL-816
GL-866-A
GL-872-A
GL-8008

G-E VACUUM CAPACITORS

GL-1L21
GL-1L22
GL-1L23
GL-1L24
GL-1L25
GL-1L31
GL-1L32
GL-1L33
GL-1L36
GL-1L38



Whatever your location, a nearby G-E tube distributor is ready to serve you . . . with Lighthouse Larry and his expert counsel always on call for YOUR benefit!

ENERAL ELECTRIC

180-JA9

Reliable Rigs use OHMITE

BROWN DEVIL RESISTORS

Sturdy, wire-wound, vitreous-enameled resistors for voltage dropping, bias units, bleeders, etc. In 5, 10, and 20-watts; values to 100,000 ohms.

FIXED RESISTORS

Resistance wire is wound over a ceramic core, permanently locked in place, insulated and protected by Ohmite vitreous enamel. In 25, 50, 100, 160, and 200-watt stock sizes; values from 1 to 250,000 ohms.

DIVIDOHM RESISTOR

You can quickly adjus these handy vitreous-enamele resistors to the exact resistanc you want, or put on taps wher ever needed for multi-tap resistors and voltage dividers. I sizes from 10 to 200 watts, t 100,000 ohms.

LITTLE DEVIL COMPOSITION RESISTORS

Tiny, molded, fixed resistors—individually marked with resistance and wattage rating—½, 1, and 2-watt sizes, ±10% tol. Also ±5% tol. 10 Ohms to 22 megohms.

DUMMY ANTENNA RESISTORS

For loading transmitters or other r.f. sources. New, rugged, vitreous-enameled units are practically non-reactive within their recommended frequency range. 100 And 250-watt sizes, 52 to 600 ohms, ±5%.

MOLDED COMPOSITION POTENTIOMETE

A high-quality, 2-watt unit wit a good margin of safety. Resistance element is solid molded—not a film. The noise leve is low and decreases with use.

CLOSE CONTROL RHEOSTATS

Insure permanently smooth, close control. Widely used in industry. All ceramic, vitreous enameled: 25, 50, 75, 100, 150, 225, 300, 500, 750, and 1000-watt sizes.

DIRECTION INDICATOR POTENTIOMETER

Compact, low cost. Used in a simple potentiometer circuit as a transmitting element to remotely indicate the position of a rotary-beam antenna.

HIGH-CURREN TAP SWITCHE

Compact, all-cerami multipoint, rotary s lectors for a-c use. Self-cleanin silver-to-silver contacts. Rate at 10, 15, 25, 50, and 100 an peres. Two or more can b mounted in tandem.

POWER LINE CHOKES

Keep r.f. currents from going out over the power line and causing interference with radio receivers. Also used to stop incoming r.f. interference. Has a ceramic core and moistureproof coating. In 5, 10, and 20 amps.

RADIO FREQUENCY CHOKES

Single-layer wound on low power-factor steatite or bakelite cores, with moistureproof coating. Seven stock sizes for all frequencies, 3 to 520 mc. Two units rated 600 ma, others rated 1000 ma.

NEV OHM'S LAV CALCULATOI

Redesigned! Figures all Ohm Law problems, including parall resistance, with one setting the slide. Also has standard slid rule scale. Send 25¢ in coin.

SEND FOR FREE CATALOG

Stock catalog lists hundreds of units, gives helpful information.

OHMITE MANUFACTURING CO.
4822 Flournoy St. Chicago 44, Il

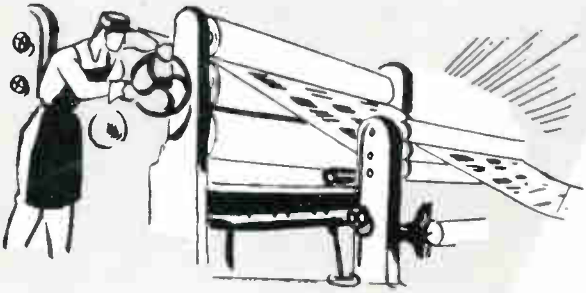
Be Right with OHMITE
Reg. U. S. Pat. O

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.

◆

The American
Radio Relay
League
Inc.

WEST
HARTFORD
7
CONN.



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, $4.25 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-six years (and thereby the oldest American radio magazine) *QST* has been the "bible" of Amateur Radio.

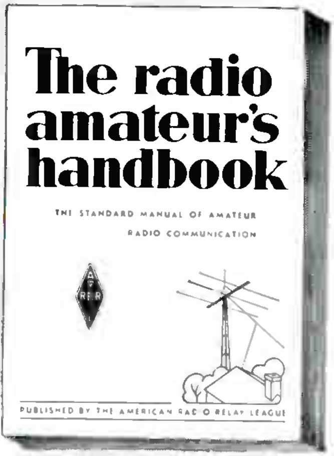
QST BINDERS

Price $2.50 each, postpaid

Available only in U. S. and Possessions

Those who take pride in the appearance of their bookshelf and wish to keep their reference file of ***QST*** in a presentable manner, appreciate the ***QST*** binder. It is stiff-covered, finished in beautiful and practical fabrikoid. Cleverly designed to take each issue as received and hold it firmly without mutilation, it permits removal of any desired issue without disturbing the rest of the file. It accommodates 12 copies of ***QST***. Opens flat at any page of any issue.

With each Binder is furnished a sheet of gold and black gummed labels for years 1940 through 1960. The proper one can be cut from the sheet and pasted in the space provided for it on the back of the binder. The back copies of ***QST*** contain the record of development of modern amateur technique. They are invaluable as technical references. Some back copies are available—list will be sent upon request.



The radio amateur's handbook

THE STANDARD MANUAL OF AMATEUR RADIO COMMUNICATION

PUBLISHED BY THE AMERICAN RADIO RELAY LEAGUE

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.

$2.50 *U.S.A. proper*

$2.75 U.S. Possessions and Canada

$3.00 Elsewhere

Buckram Bound Edition

$4.00 U.S.A. proper

$4.50

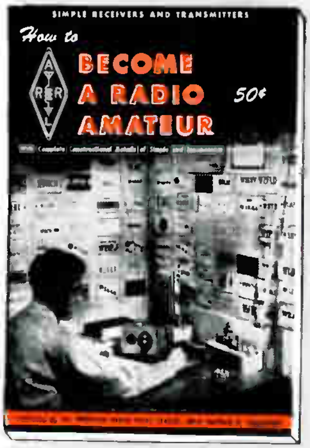
Elsewhere

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.

50c

Postpaid

Anywhere

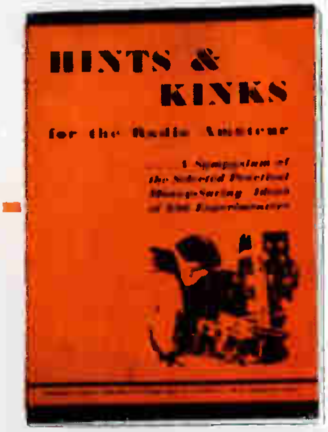


SIMPLE RECEIVERS AND TRANSMITTERS

How to

BECOME A RADIO AMATEUR

50¢



HINTS & KINKS

for the Radio Amateur

Amateurs are noted for their ingenuity in overcoming 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 procedures for war surplus. It will return its cost many times in money savings—and it will save hours of time.

$1.00 *Postpaid*
U.S.A. and Possessions,
and Canada
Elsewhere
$1.25

A comprehensive manual of antenna design and construction, by the headquarters staff of the American Radio Relay League. Sixteen chapters, 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.

$1.00 *Postpaid*
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and Canada
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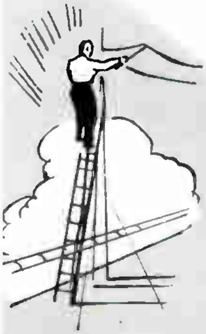
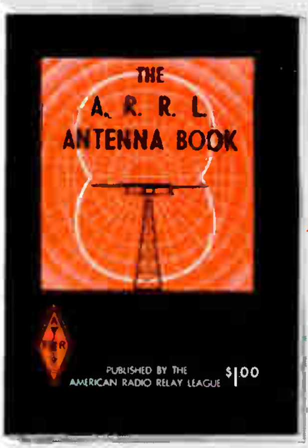
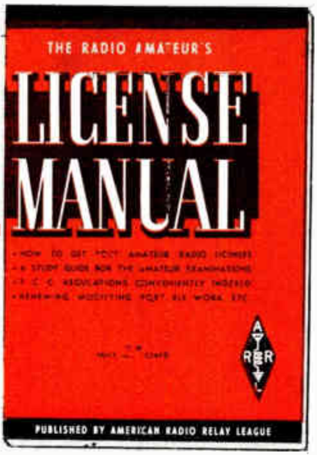
$1.25





THE
A. R. R. L.
ANTENNA BOOK

PUBLISHED BY THE
AMERICAN RADIO RELAY LEAGUE
$1.00



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.

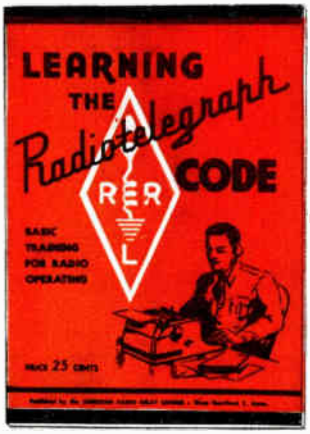
25c

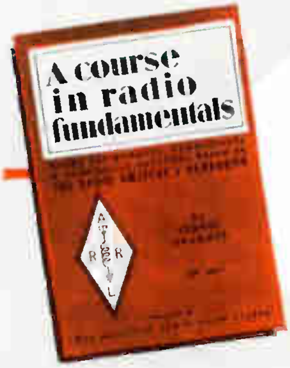
Postpaid

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.

25c

Postpaid





"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

Also available in loose-leaf form, punched for a three-ring binder, 100 8½ x 11 log Sheets, **75c.**

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.

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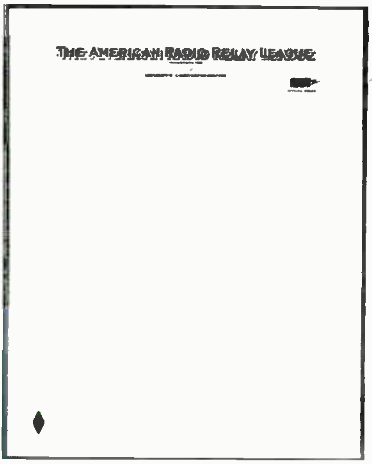
MEMBERSHIP



THE AMERICAN RADIO RELAY LEAGUE

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.

$1.00 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 1C 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.*

To Handbook readers who are not ARRL members . . .

For thirty-seven years the American Radio Relay League has been the organized body of amateur radio, its representative in this country and abroad, its champion against attack by other interests, its leader in technical progress, its center of operating activities.

Join the League

ARRL is an organization that *does things*. The League protects amateur interests in domestic legislation and regulations and at international conferences. It stages annual operating events such as the Sweepstakes, Member Party, Field Day and DX contests; offers appointments such as Official Experimental Station, Emergency Coordinator, and Official Bulletin Station; and issues awards for achievement in operating skill such as Worked-All-States, DX Century Club, and Code Proficiency. It handles foreign QSL cards for you, answers your technical and regulatory questions, provides you or your clubs with training aids and operating literature — to name just a few of its many services.

As an ARRL member, you will be posted on amateur affairs. *QST* will be delivered to your door each month, chock full of the latest news of ham doings, not to mention a wealth of technical and constructional material on amateur gear.

We need you in this big organization of radio amateurs. You need the League and its services to get the most out of your amateur activities.

A bona-fide interest in amateur radio is the only essential requirement for membership, but full voting membership is granted only to licensed amateurs of the United States and Canada.

An application blank for League membership and *QST* subscription is printed on the following page.

The **American Radio Relay League, *Inc.***

Headquarters: WEST HARTFORD, CONNECTICUT, U. S. A.

Application for Membership

AMERICAN RADIO RELAY LEAGUE

Administrative Headquarters: West Hartford, Conn., U. S. A.

. . . . .19. . . .

AMERICAN RADIO RELAY LEAGUE,
West Hartford, Conn., U. S. A.

Being genuinely interested in Amateur Radio, I hereby apply for membership in the American Radio Relay League, and enclose $4.00* in payment of one year's dues, $2.00 of which is for a subscription to QST for the same period. Please begin my subscription with the. . . . . .issue.

The call of my station is. . . . .

The class of my operator's license is. . . . .

I belong to the following radio societies. . . . .

. . . .

Send my Certificate of Membership ☐ or Membership Card ☐ (Indicate which) to the address below:

Name. . . . .

. . . .

. . . .

A bona fide interest in amateur radio is the only essential requirement but full voting membership is granted only to licensed radio amateurs of the United States and Canada. Therefore, if you have a license, please be sure to indicate it above.

(*$4.00 in the United States and Possessions, $4.25, U. S. funds, in Canada, $5.00, U. S. funds, in all other countries.)

COMPONENTS FOR EVERY APPLICATION


LINEAR STANDARD
High Fidelity Ideal


HIPERM ALLOY
High Fidelity . . . Compact


ULTRA COMPACT
Portable . . . High Fidelity


OUNCER
Wide Range . . . 1 ounce


SUB OUNCER
Weight ⅓ ounce


COMMERCIAL GRADE
Industrial Dependability


SPECIAL SERIES
Quality for the "Ham"


POWER COMPONENTS
Rugged . . . Dependable


VARITRAN
Voltage Adjustors


MODULATION UNITS
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 ½ — 10,000 cyc.


EQUALIZERS
Broadcast & Sound


PULSE TRANSFORMERS
For all Services


SATURABLE REACTORS
Power or Phase Control


LARGE UNITS
To 100KW Broadcast


PLUG-IN TYPE
Quick change service


CABLE TYPE
For mike cable line


VERTICAL SHELLS
Husky . . . Inexpensive


REPLACEMENT
Universal Mounting


STEP-DOWN
Up to 2500W . . . Stack

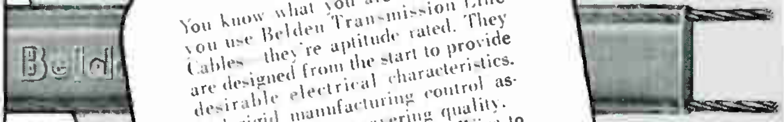

LINE ADJUSTORS
Match any line voltage


CHANNEL FRAME
Simple Low cost

United Transformer Co.

150 VARICK STREET NEW YORK 13, N. Y.
EXPORT DIVISION 13 EAST 40th STREET, NEW YORK 16, N. Y., CABLES: "ARLAB"

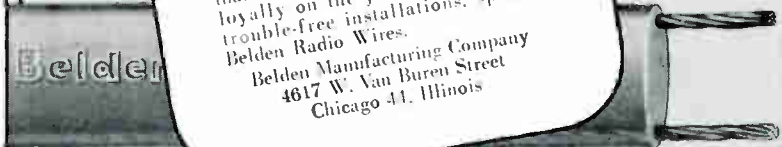
Be Sure of Your Installations –
Get the *Aptitude-Tested* TRANSMISSION LINE



APTITUDE RATING **No. 8225**

Frequency (mc)	Attenuation per 100 ft
100.	1.1
200.	1.7
300.	2.2
400.	2.7

For use with television and FM receiving antenna. Exceptionally low losses at high frequencies.



APTITUDE RATING **No. 8235**

Frequency (mc)	Attenuation per 100 ft
100.	1.10
200.	1.73
300.	2.28
400.	2.74

For use with television and FM receiving antenna; also for low-power transmitting antenna.

You know what you are doing when you use Belden Transmission Line Cables—they're aptitude rated. They are designed from the start to provide desirable electrical characteristics, and rigid manufacturing control assures constant, unwavering quality.

You can safely put Belden Wire to work for you, and know for sure how it will perform. You can know, too, that it will have the stamina to stay loyally on the job for years. For trouble-free installations, specify Belden Radio Wires.

Belden Manufacturing Company
4617 W. Van Buren Street
Chicago 44, Illinois

APTITUDE RATING **No. 8227**

Frequency (mc)	Attenuation per 100 ft
100.	4.1
200.	6.4
300.	8.4
400.	10.2

For use with television and FM antenna in extremely noisy locations.

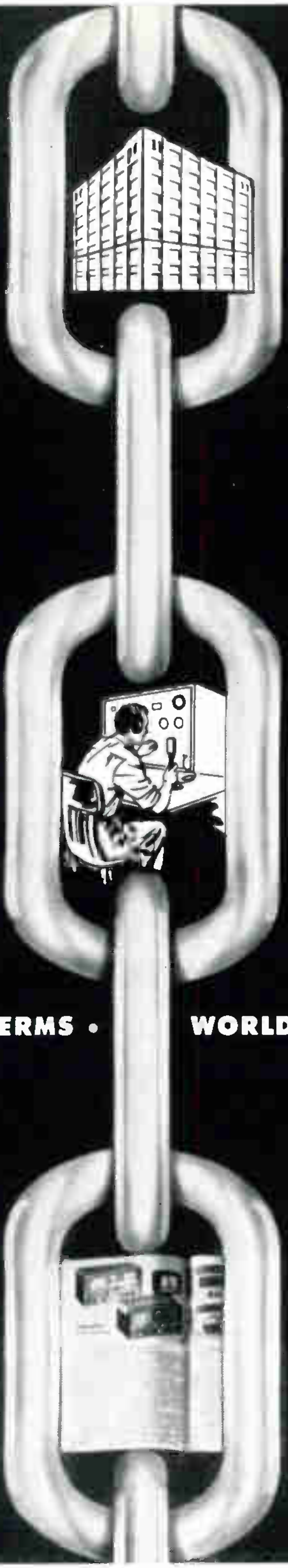
APTITUDE RATING **No. 8240**

Frequency (mc)	Attenuation per 100 ft
100.	4.10
200.	6.20
300.	8.00
400.	9.50

For use with radio frequency transmission video, test equipment, and pulse transmission.

Belden *Radio* **WIRE**

The *Aptitude-Tested* LINE



IT'S ALLIED

FOR

COMPLETE
QUICK
DEPENDABLE

AMATEUR

STATION SUPPLY SERVICE

EASIEST TERMS • WORLD'S LARGEST STOCKS • BEST VALUES

Depend on ALLIED to bring you the products of each and every one of the dependable manufacturers represented in the Handbook. We carry the world's largest stocks of Amateur station supplies—and, of course, we can supply all the components you'll need to build any circuit described in this or any other publication . . . You can always depend on ALLIED to give you every buying advantage.

Write for your FREE copy of the latest ALLIED Catalog. It's the only complete Amateur Buying Guide to all your station supply needs. Keep it handy!

ALLIED RADIO

833 W. Jackson Blvd., Chicago 7, Ill.

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 **$7.65**, **No. 58**—200 watt **$9.85**, **No. 78**—300 watt **$12.05** and **No. 98**—550 watt **$14.25**. 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.

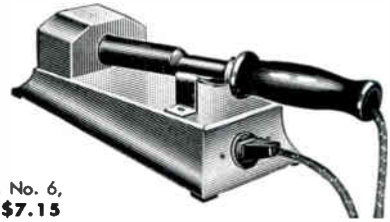
The No. 61 is regularly wound to 25 watts at 105–120 volts, but may be had in higher wattages, when required in quantities at no extra cost. List price of iron is $5.45. Tips A, B, or C 40c each list. Irons are available thru any of the better tool or Radio & Electronic Supply houses. If your distributor 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	**$4.95**
Model No. 36—2¼ lbs. cap. 250 watt. Net	**$6.05**
Model No. 60—3¾ lbs. cap. 325 watt. Net	**$7.15**

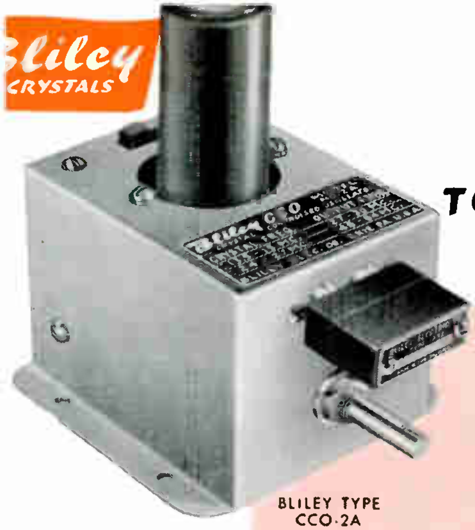
• Temperature Control Stand



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............................**$7.15**

ELECTRIC SOLDERING IRON CO., Inc.

2551 WEST ELM STREET **DEEP RIVER, CONN., U. S. A.**



Bliley CRYSTALS

TOP CHOICE FOR TWENTY YEARS!

BLILEY TYPE
AX2

BLILEY TYPE
AX3

BLILEY TYPE
CCO-2A

ONLY Bliley offers the famous CCO-2A rystal controlled oscillator for 2-6-10-11 meters. Direct output on 6-10-11 meters. Use Bliley type AX2 20 meter crystal for top performance on 10-11 meters; Bliley type AX3 for 6 and 2 meter operation. The ideal nucleus for your new equipment.

BLILEY TYPE AX2

Bliley Type	Supplied Tolerance (kc)	Range (kc)
AX2	±1	1803 - 1822 1878 - 1897 1903 - 1922 1978 - 1997
AX2	±2	3500 - 3999
AX2	±2	7000 - 7425
AX2	±30	12500 - 13500
AX2	±30	13480 - 13615
AX2	±30	14000 - 14850

Calibrated to ± .002% Drift less than .0002% per °C.

BLILEY TYPE AX3

Bliley Type	Supplied Tolerance (kc)	Range (kc)
AX3	±5	24000 - 24333
AX3	±5	25000 - 25500

Specially designed third overtone crystal produced for the Bliley CCO-2A. Calibrated to ± .003%. Drift less than .0002% per °C.

BLILEY TYPE
BH6

COMMERCIAL designation for HC-6/U holder; metal-to-glass hermetic seal. Supplied as types CR-18; CR-23; CR-27; CR-32 for Military and CAA specifications. Frequency range 1-100 mc.

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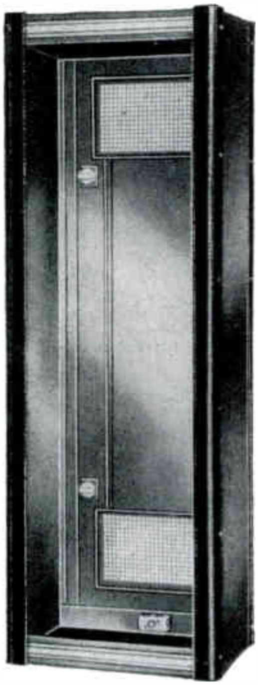

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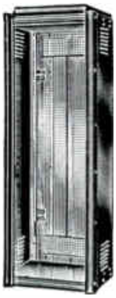

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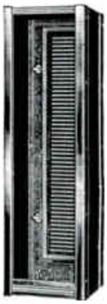

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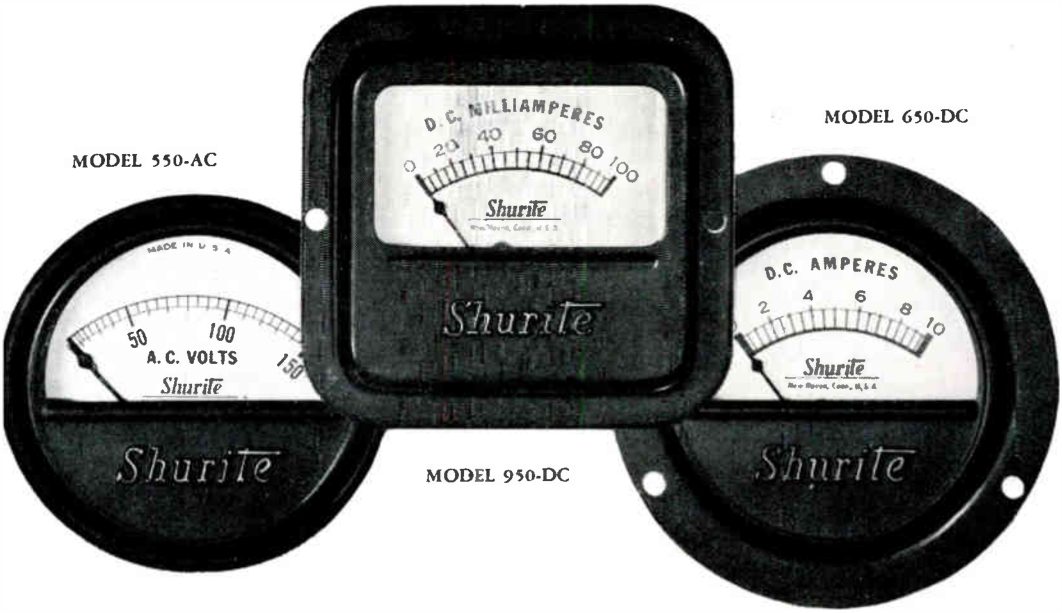

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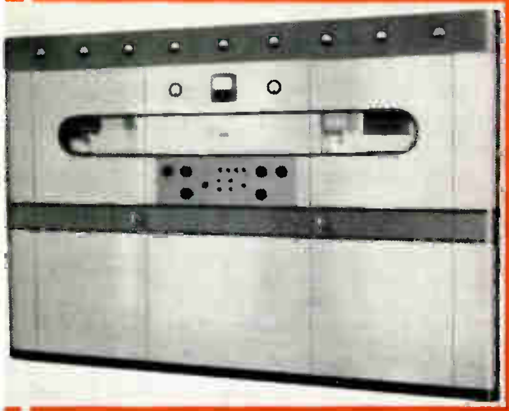

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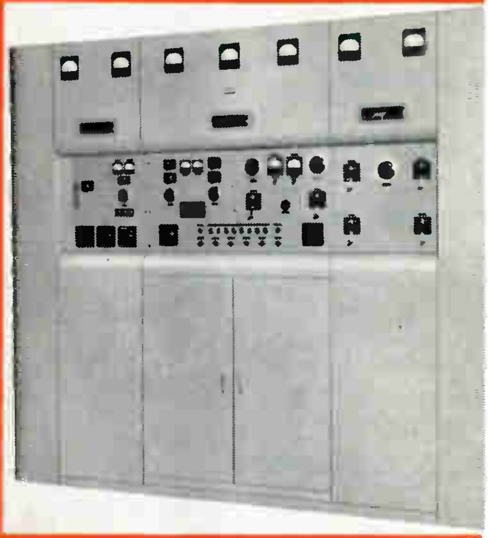

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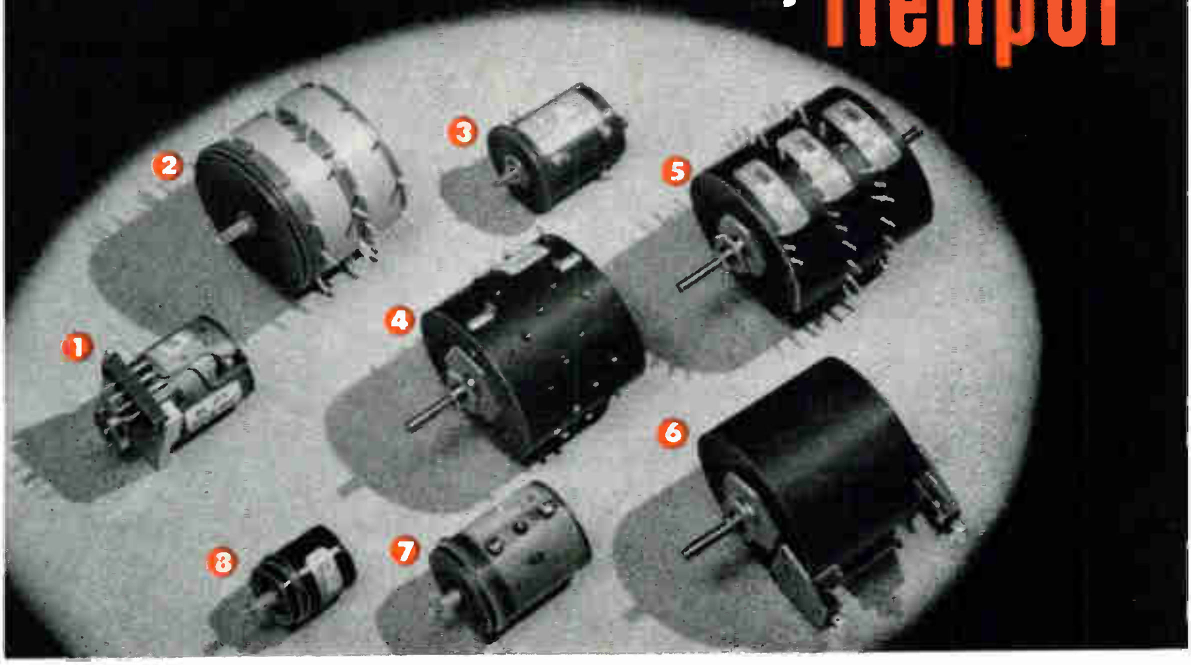

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② *This four-gang assembly of Model F single-turn potentiometers has a special machined aluminum front end for servo-type panel mounting, with shaft supported by precision ball bearings and having a splined and threaded front extension. Each of the four resistance elements contains 10 equi-spaced tap connections with terminals, and all parts are machined for greatest possible stability and accuracy.*

③ This standard Model A, 10-turn HELIPOT has been modified to incorporate ball bearings on the shaft and a special flange (or ring-type) mounting surface in place of the customary threaded bushing. This HELIPOT also contains additional taps and terminals at the ¼- and 9¾-turn positions.

④ *This standard Model B, 15-turn HELIPOT has a total of 40 special tap connections which are located in accordance with a schedule of positions required by the user to permit external resistance padding which changes the normally-linear resistance vs. rotation curve to one having predetermined non-linear characteristics. All taps are permanently spot-welded and short out only one or two turns on the resistance element—a unique HELIPOT feature!*

⑤ This six-gang assembly of standard Model F single-turn potentiometers has the customary threaded bushing mountings, and has shaft extensions at each end. The two center potentiometers each have 19 equi-spaced, spot-welded tap connections brought out to terminals. *Each tap shorts only two turns of .009" diameter wire on the resistance element.*

⑥ *This Model B, 15-turn HELIPOT has been modified to incorporate, at the extreme ends of mechanical and electrical rotation, switches which control circuits entirely separate from the HELIPOT coil or its slider contact.*

⑦ This 10-turn HELIPOT has many design features similar to those described for unit No. 1, plus the following additional features . . . a servo-type front end mounting . . . splined and threaded shaft extension . . . and a center tap on the coil. All components are machined to the highest accuracy, with concentricities and alignments held in some places to a few *ten-thousandths* of an inch to conform to the precision of the mechanical systems in which this HELIPOT is used. *Linearity accuracies frequently run as high as ±0.010%!*

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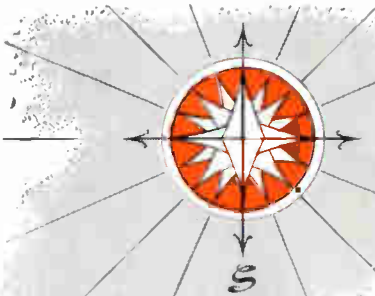

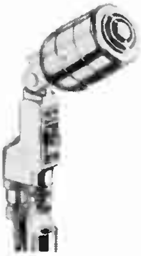

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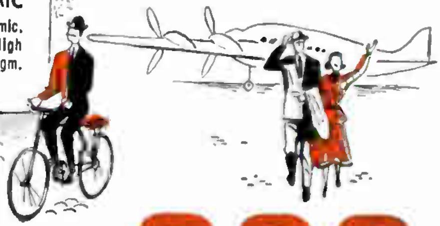

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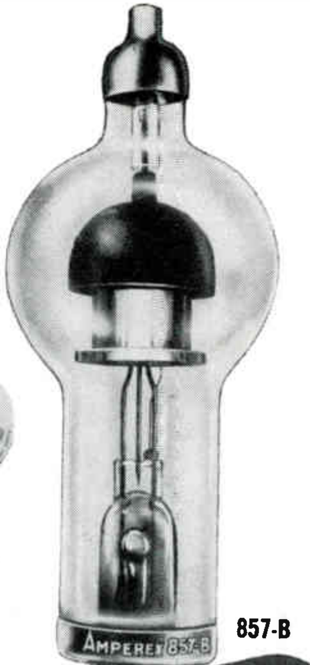

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			CCS	ICAS
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PT8312	1200-0-1200	1000	325	405
PT8313	1475-0-1475	1250	250	310
PT8314	1790-0-1790	1500	225	280
PT8315	2065-0-2065	1750	200	250

DC output rated CCS at load terminals of single-section reactor-input filter, ICAS with single-section capacitor-input filter. Primaries for 117 volts, 60 cycles.

TYPE & PART NO.	SECONDARY AC VOLTS	DC OUTPUT VOLTS	DC OUTPUT MA.	TYPE FILTER	RECTIFIER
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PC8302	515-0-515	385 500	235 200	Reactor Input Capacitor Input	5U4G 5R4GY
PC8303	665-0-665	500 750	250 200	Reactor Input Capacitor Input	5R4GY 5R4GY
PC8304	760-0-760	600 800	265 200	Reactor Input Capacitor Input	2-5R4GY 5R4GY
PC8305	920-0-920	750 1000	250 200	Reactor Input Capacitor Input	2-5R4GY 5R4GY
PC8306*	920-0-920	750 1000	150 150	Reactor Input Capacitor Input	5R4GY 5R4GY
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Here's your chance to learn code quickly and easily . . . to get maximum speed and proficiency in code operation . . . right in your own home. The forty-year record of the CANDLER SYSTEM in training beginners to develop and apply their talents to top speed and efficiency is sound proof of what the CANDLER SYSTEM *offers you*

Think of the thrills and excitement you can have communicating by code with people all over the world . . . when you become a good Amateur Operator. Or, you may have your sights set on becoming a highly-paid, expert Commercial Operator. In either case, you can *save time and money* learning the fundamentals of code sending and receiving and the principles of fast, efficient operating with the world-famous CANDLER SYSTEM.

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Are you hoping for a nice promotion . . . a bigger salary . . . as an expert Operator? Then, now is the time to improve your speed and skill in code sending and receiving. The CANDLER SYSTEM OF CHAMPIONS shows you how to improve your code technique — increase your speed — develop your skill to the highest point — eliminate nervous tension — and pave the way to a top-bracket position as an expert Operator.

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The "REX" Crystal Microphone is a striking-looking, low-cost hand-held microphone. Fits snugly in the hand, sits firmly on a desk, or can be placed on a stand. Recommended where good quality speech reproduction is required and low-cost is an important factor. With stand adapter.

Model 710A . . . . List Price $10.95
Model 710S . (with switch) . List Price $12.95

The "HERCULES" Controlled Reluctance Microphone provides clear reproduction, high speech intelligibility, high output and ruggedness *at an amazingly low price!* Can be used indoors or outdoors—fits in the hand, sits firmly on a desk, can be placed on a stand. (With stand adapter.)

Model 510C . . . . List Price $14.50
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The "DISPATCHER" is a Controlled Reluctance Microphone dispatching unit that handles the most severe field requirements of amateur rigs, paging and dispatching systems. Supplied with 3-conductor shielded cable, and is wired to operate both microphone and relay circuits.

Model 520SL-7 (7' cable) **. . List Price $35.00**
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The "SONODYNE", a high-output dynamic microphone with wide-range frequency response. Has moving coil unit. Features a Multi-Impedance switch. A rugged unit with high sensitivity, yet perfect for Hams in high temperature and high humidity locations.

Model 51 . . . . List Price $42.50

The multi-impedance "UNIDYNE" is a high quality Super-Cardioid Dynamic Microphone for hams whose rigs must provide dependable performance even under difficult conditions. Reduces random noise pickup by 73%.

Model 55 . . . . List Price $69.50

The "MONOPLEX", the only super-cardioid crystal microphone. Has high-output, wide-range frequency response. Perfect for Hams who want the extra "push" that insures a "strong" voice. Features high-quality performance at low cost.

Model 737A . . . . List Price $39.75

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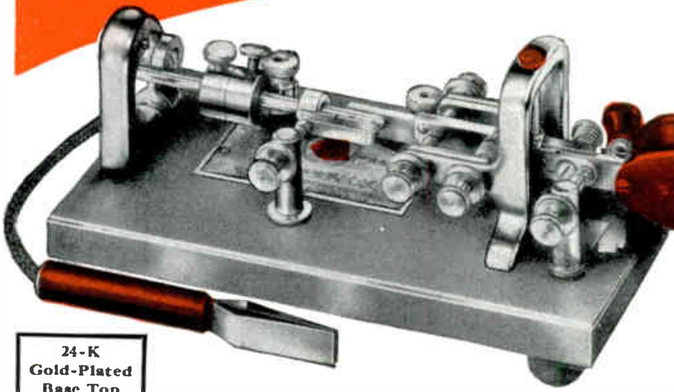

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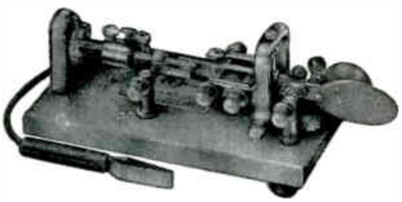
SENSATIONAL NEW SUPER DE LUXE VIBROPLEX

The PRESENTATION — No ifs and ands about it, this is the world's outstanding key! Created for those who want the best! This is it! The most beautiful, also the easiest to use! Made to fine instrument standards, it has many exclusive improvements that make sending easy for everyone, including Patented Jewel movement for smoother, easier operation and longer life; Super-Speed Control Main Spring providing a range of speed from dead slow to lightning fast; Touch Control that permits instant adjustment of key tension to individual touch and style of sending. Until you've used this key you've no idea how easy sending can be.

Jewel-like 24-K gold-plated base top. Polished chromium parts. Colorful red switch knob, finger and thumb-piece. Weight 3 lbs. 8 oz. With cord and wedge ready for use, **$29.95.**



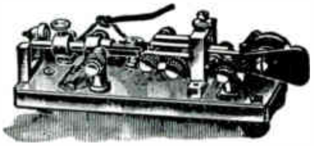
The ORIGINAL, above — This famous key has won an international reputation for signal perfection, speed and operating ease by thousands of the world's finest operators! A true professional key precision machined for superior operating performance. Built for hard usage and long life, with a range of speed to meet every operating need. Black crystal base. Polished chromium parts. Weight 3 lbs. 8 oz., **$17.95.**

De Luxe — Handsome polished chromium base and parts. Red trim. Jewel movement. Cord and wedge, **$22.50.**

The LIGHTNING BUG — New design, beautifully finished and moderately priced! Precision machined for TOP performance make it one of the easiest-working keys available. Has a flat vibrating lever . . . instantly adjustable dot contact spring . . . bridged damper frame and many other features contributing to smooth, easy operating. Maintains a clear snappy signal at all operating speeds. A popular choice. Black crystal base. Polished chromium parts. Weight 3 lbs. 8 oz., **$15.95.**

De Luxe — Smart polished chromium base and parts. Red trim. Jewel movement. Cord and wedge, **$21.50.**

The CHAMPION — Full-size! Nothing like it anywhere near this low price! Patterned after the Lightning Bug and maintaining much of the smoothness and handling ease of that popular key! Responds to the lightest touch. An especially fine key for the beginning amateur because it is so simple and easy to use. Don't let the low price mislead you! This key will give you years of sending enjoyment. Black crystal base. Chrome-plated parts. Weight 3 lbs. 8 oz. Less circuit closer, cord and wedge. For radio use only, **$12.95.**



The BLUE RACER — An unusually compact key built extra sturdy like the Original and capable of the same operating ease and excellence for which that key is so well known. A space saver . . . only about half the size of the Original, with a range of speed to meet every operating need. Ideally suited to radio requirements. Black crystal base. Polished chromium parts. Weight 2 lbs. 8 oz., **$17.95.**

De Luxe — Smart polished chromium base and parts. Red trim. Jewel movement. Cord and wedge, **$22.50.**

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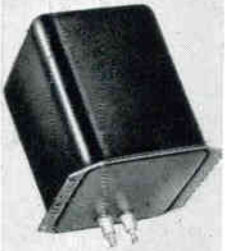

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Here they are—top-performing plate transformers and matching filter reactors. They're conservatively designed, with ample insulation throughout. They operate with a temperature rise of 40° to 50°C at full load, 60 cycles, under CCS duty. Under ICAS conditions, duty cycle is 15 min. on and 15 off, 40°-50° temperature rise.

Catalog No.	Max. Pri. VA	Secondary A-C Load Volts	D-C Volts after filter	D-C Ma. CCS	ICAS
P-45	185	675-0-675 575-0-575	400 500	250	325
P-67	250	900-0-900 735-0-735	750 600	250	325
P-107	310	1150-0-1150 870-0-870	1000 750	250	350
P-1240	360	1425-0-1425* 600-0-600	1250 400	150 200	200 260
P-1512	550	1710-0-1710 1430-0-1430	1500 1250	300	425
P-2520	915	2820-0-2820 2260-0-2260	2500 2000	300	425
P-3025	1850	3450-0-3450 2850-0-2850	3000 2500	500	700

*Both secondaries may be rectified simultaneously

FILTER REACTORS

Catalog No.	Inductance in Henries	Max. D-C Ma.	D-C Resistance, Ohms	Insulation Volts RMS
R-67	6	700	35	10,000
R-105	10	500	40	9,000
R-65	6	500	35	9,000
R-103	10	300	40	7,500
R-63	6	300	35	7,500

Filament Transformers

CHICAGO offers a complete line of filament transformers, rated to provide voltages and currents for heating a wide range of receiving and transmitting tubes. Filament units geared to today's Amateur circuits are fully described in our New Equipment Catalog. Write for your FREE copy today.

Splatter Chokes

Designed specifically for use in high level "clipper" filters. They eliminate splatter caused by heavy modulation, limit band width to 3,000 cycles, effectively increase "get through" phone signal ability, prevent negative peak clipping. Windings tapped to cover inductance range from .02 to 1.5 hys at relatively constant Q. Adequately insulated to withstand high peak voltages.

Catalog No.	D-C Ma.	Insulation Volts	Inductance in Henries
SR-300	300	7,500	.02 to 1.5 Hy
SR-500	500	10,000	.02 to 1.5 Hy

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Low Pass Filter No. LPF-1

Designed to provide optimum attenuation of audio frequencies above 3,000 cycles. Economically and effectively confines a speech signal within narrow frequency limits. The filter operates out of a source impedance of 15,000 ohms (as presented by plate of single 6C5, 6J5, or triode-connected 6J7) into a 100,000 ohm grid. Step-up ratio of 1: 2.63 is obtained when primary is connected to 100,000 ohm secondary. A plate blocking condenser is required to prevent D-C from saturating the coils in the filter proper. Filter operates satisfactorily at signal levels of +10 volts RMS across the 100,000 ohm secondary, or of +0.7 volts RMS at 500 ohms.

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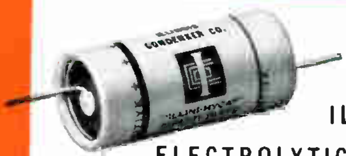

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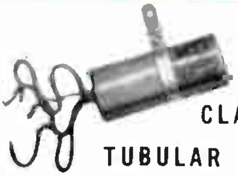

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Built for long life under severest operating and climatic conditions. Leads color coded and securely anchored. Common negative or multiple negative units for all service applications. Available in low voltage common negative, separate negative 4 leads, multiple units, high voltage — single units, high surge — single units, and high voltage, multiple units. Working voltages DC from 3 to 600.

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Offer a wider range of voltage and capacity types than heretofore possible in units of comparable size. Characteristics superb. Capacities always plus, low power factor, low leakage. Hermetically sealed in seamless drawn aluminum cans. Soldering lugs heavily tinned.

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(Inverted Screw Mounting)

Built to operate under severest conditions. Units completely sealed in inner impregnated tube then resealed. Design permits maximum heat dissipation, higher temperature and voltage surges. Available in regular, dual negative 4 leads, triple and quadruple units common negative sections. Single and dual units 450 to 600 DCWV.

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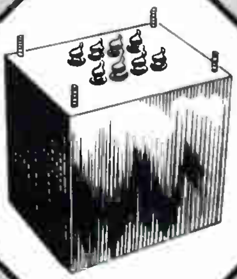

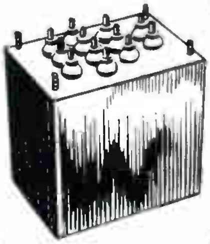

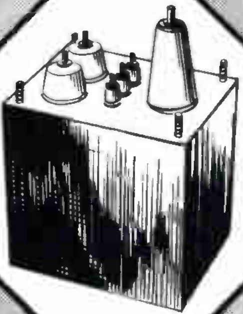

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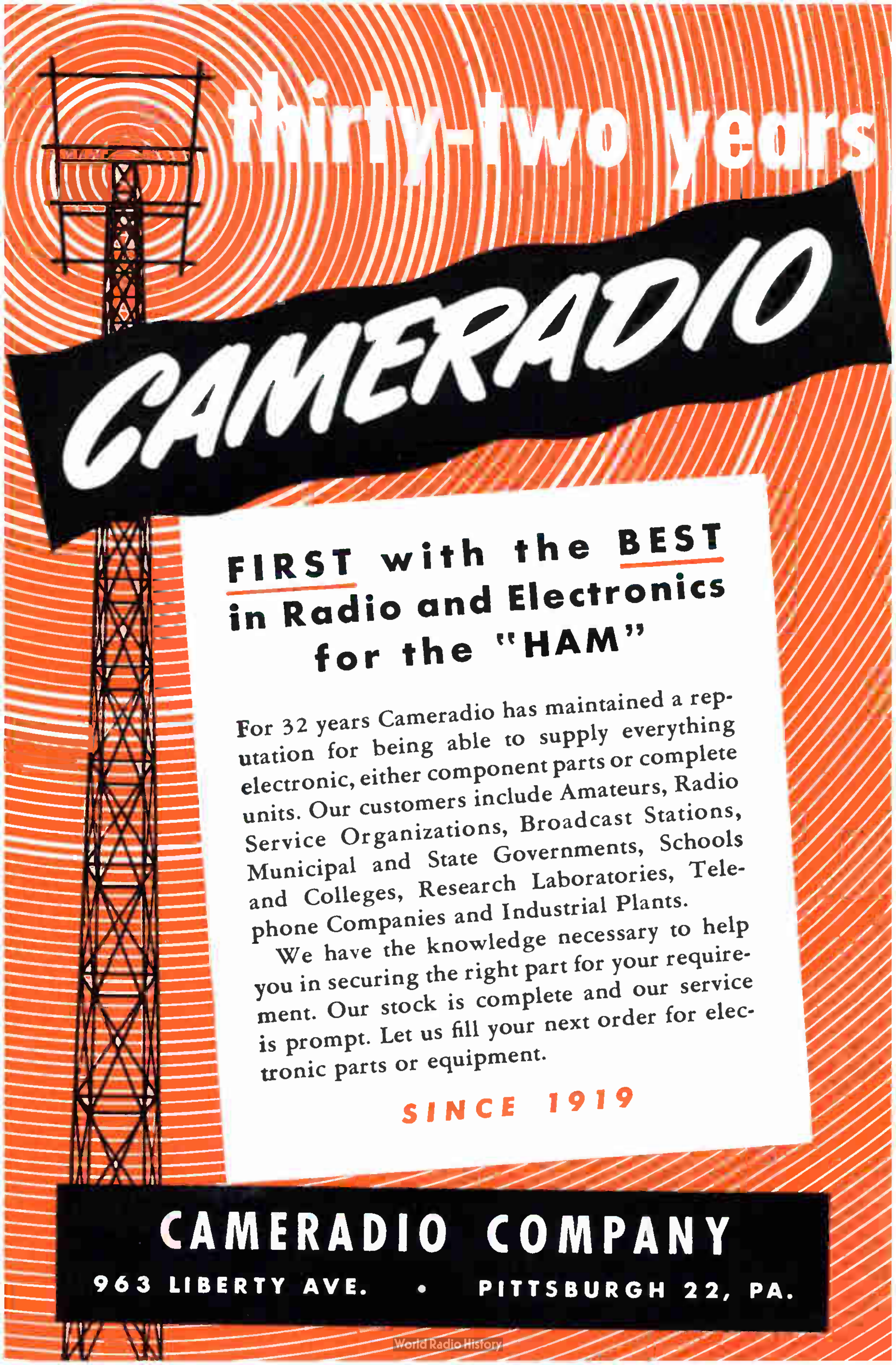

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Laboratory-precision VTVM for trigger-fast operation and lifetime service. 15 different ranges. Large 4½" meter, can't-burn-out circuit. New zero center for TV & FM discriminator alignment. Electronic AC & DC ranges: 0-5, 10, 100, 500, 1000 v. (30,000 volts & 200 MC with HVP-1 & P-75 probes). Ohmmeter ranges, .2 ohms to 1000 megs. DB scale. New stable double-triode balanced bridge circuit—extreme accuracy. 26 megs DC input impedance. 3-color etched rubproof panel; steel case. 115 v., 60 cycle AC. 9-7/16 x 6 x 5".

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For FM-AM precision alignment and TV marker frequencies. Vernier Tuning Condenser. Highly stable RF oscillator, range: 150 KC—102 MC with fundamentals to 34 MC. Separate audio oscillator supplies 400-cycle pure sine wave voltage. Pure RF, modulated RF or pure AF for external testing. 115 v., 60 cycle AC. 10 x 8 x 4¾".

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New DELUXE SIGNAL GENERATOR

A laboratory-precision generator EICO Service-Engineered with 1% accuracy. Extremely stable, frequency 75 KC—150 MC in 7 calibrated ranges. Illuminated hairline vernier tuning. VR stabilized line supply. 400-cycle pure sine wave with less than 5% distortion. Tube complement: 6X5, 7F7, 6C4, VR-150. 3-color etched panel; rugged steel case. 115 v., 60 cycle AC. 12 x 13 x 7".

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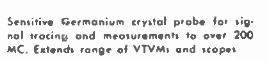

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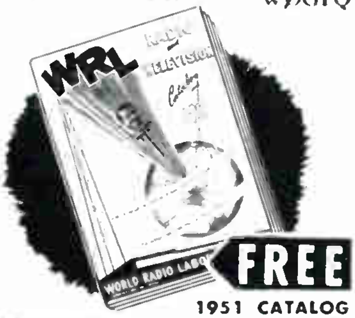

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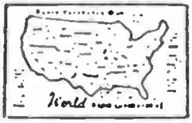

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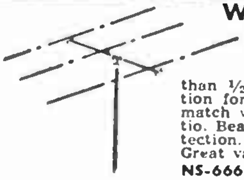

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ULL WAVE RECTIFIER EL 1C

C. Output (Amps.) ..	1.0
eak Anode Current ..	4.0
Peak Inverse Volts	725
Filament Volts	2.5
Filament Amperes	6.0
Overall Length	5½"

FULL WAVE RECTIFIER EL 3C

D.C. Output (Amps.) ..	2.5
Peak Anode Current ..	10.0
Peak Inverse Volts	725
Filament Volts	2.5
Filament Amperes	11.5
Overall Length	7"

FULL WAVE RECTIFIER EL 6C

D.C. Output (Amps.) ..	6.4
Peak Anode Current ..	25.6
Peak Inverse Volts	725
Filament Volts	2.5
Filament Amperes	17.0
Overall Length	7½"

HALF WAVE RECTIFIER EL 6B & EL 6F

D.C. Output (Amps.) ..	6.4
Peak Anode Current ..	40.0
Peak Inverse Volts	920
Filament Volts	2.5
Filament Amperes	21
Overall Length (6B) ..	9"
Overall Length (6F) ..	8¼"

(Panel Mounting)

HALF WAVE RECTIFIER EL 16F

D.C. Output (Amps.) ..	16.0
Peak Anode Current ..	96.0
Peak Inverse Volts	620
Filament Volts	2.5
Filament Amperes	36
Overall Length	15⅝"

(Panel Mounting)

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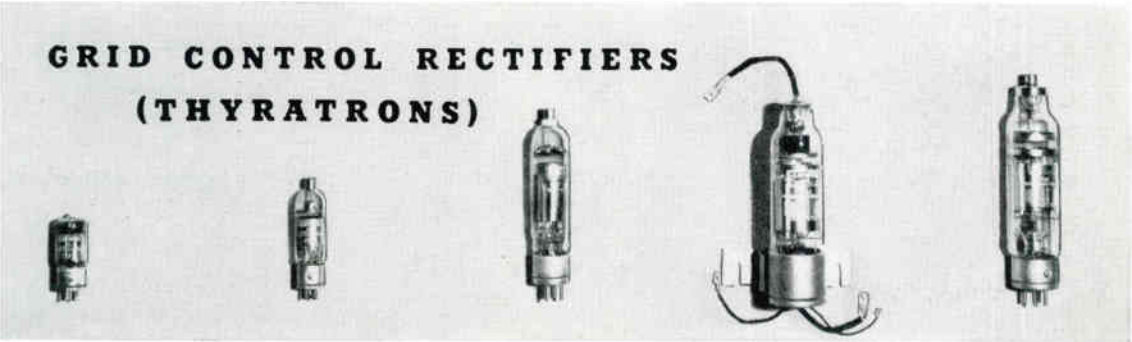


EL C1K

C. Output (Amps.) ..	1.0
eak Anode Current ..	8.0
eak Forward Volts	1000
eak Inverse Volts	1250
ilament Volts	2.5
ilament Amperes	6.3
verall Length	4¼"

EL C3J

D.C. Output (Amps.) ..	2.5
Peak Anode Current ..	30.0
Peak Forward Volts	750
Peak Inverse Volts	1250
Filament Volts	2.5
Filament Amperes	9.0
Overall Length	6⅛"

EL C6J

D.C. Output (Amps.) ..	6.4
Peak Anode Current ..	77.0
Peak Forward Volts	750
Peak Inverse Volts	1250
Filament Volts	2.5
Filament Amperes	21.0
Overall Length	9"

EL C16J

D.C. Output (Amps.) ..	16.0
Peak Anode Current	160.0
Peak Forward Volts	1000
Peak Inverse Volts	1250
Filament Volts	2.5
Filament Amperes	31.0
Overall Length	10"

(Panel Mounting)

EL C6C

D.C. Output (Amps.) ..	6.4
Peak Anode Current ..	77.0
Peak Forward Volts	2000
Peak Inverse Volts	4000
Filament Volts	2.5
Filament Amperes	24.0
Overall Length	11"

EL C1B/A

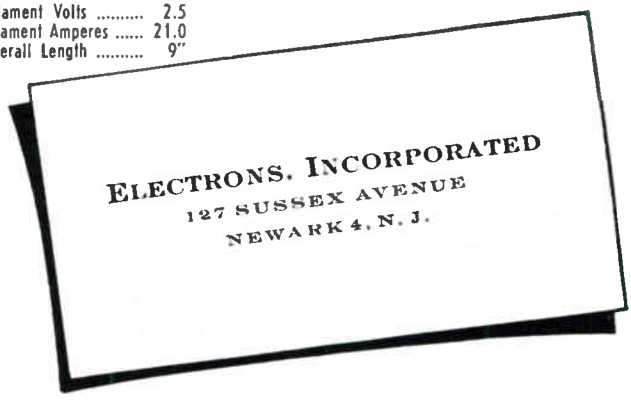
C. Output (Amps.) ..	1.0
eak Anode Current ..	8.0
eak Forward Volts	750
eak Inverse Volts	1250
ilament Volts	2.5
ilament Amperes	6.3
verall Length	4½"

EL C3J/A

D.C. Output (Amps.) ..	2.5
Peak Anode Current ..	30.0
Peak Forward Volts	1000
Peak Inverse Volts	1250
Filament Volts	2.5
Filament Amperes	9.0
Overall Length	6⅛"

EL C6J/A

D.C. Output (Amps.) ..	6.4
Peak Anode Current ..	77.0
Peak Forward Volts	1000
Peak Inverse Volts	1250
Filament Volts	2.5
Filament Amperes	21.0
Overall Length	9"



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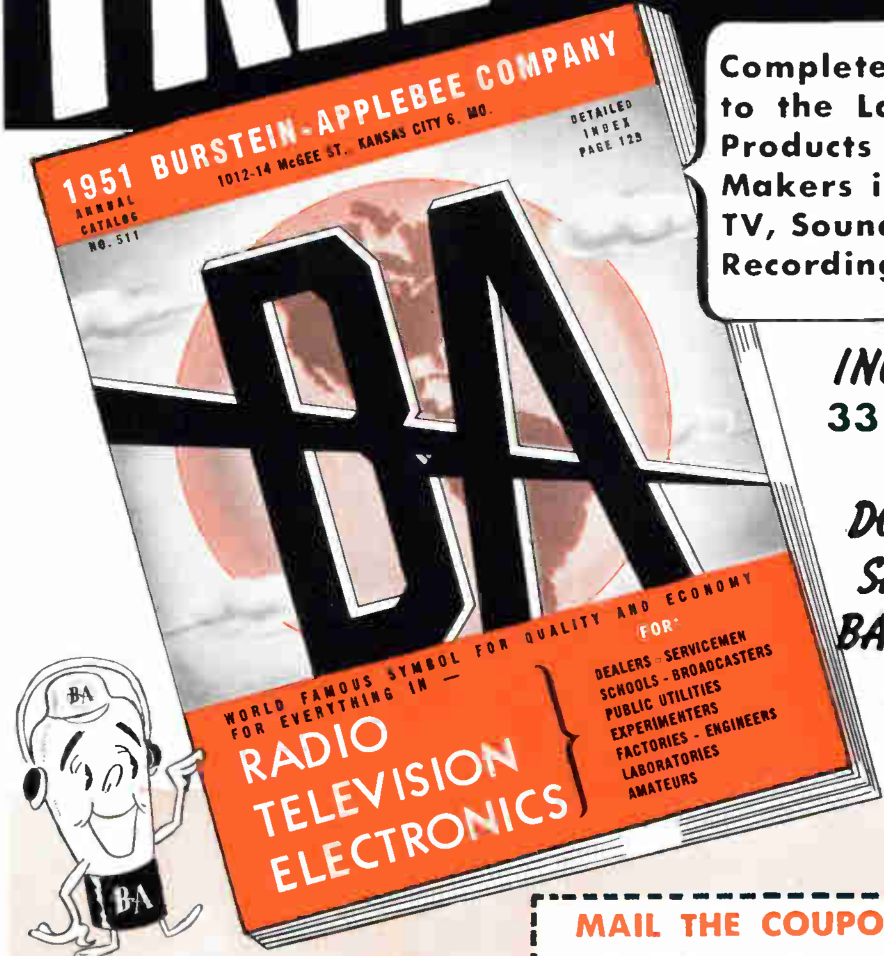

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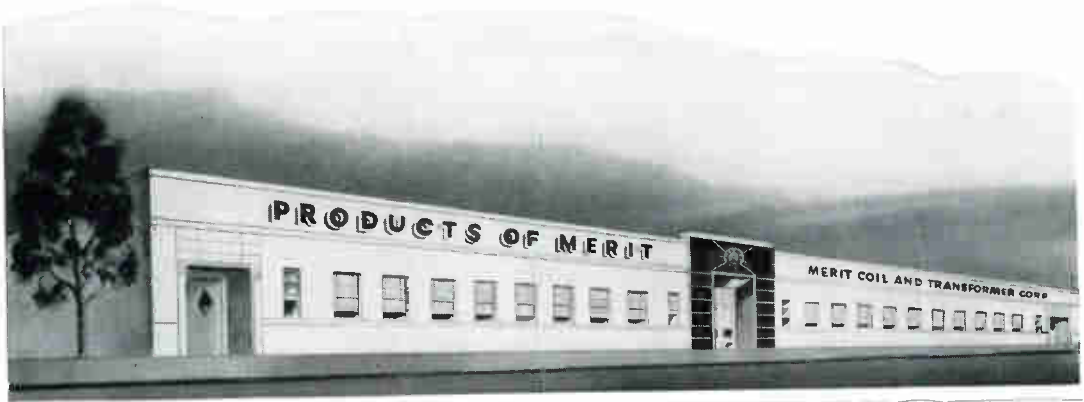

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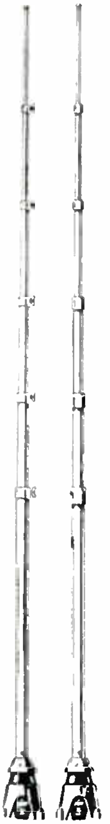
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No.	*Description*	*Length Extended*	*Length Collapsed*	*O.D. Base*	*I.D. Base*	*Weight each, lbs.*
112-M	2-sec. telescop'g	11′8″	6′1″	.656″	.556″	4
318-M	3-sec. telescop'g	17′3″	6′2″	.875″	.775″	7
224-M	4-sec. telescop'g	22′9″	6′3″	1.063″	.963″	11
130-M	5-sec. telescop'g	28′3″	6′4″	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′4″	.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

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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.

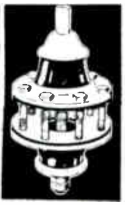
PREMAX MOUNTINGS AND INSULATORS


Type 1


Type 1X


Type 1H


Type 2


Type 6


Type 10-S


Type 8-C


Type 3


Type 13-S


Type 13H


Type 9-C


Type 10-C


Type 7

Type 1 Base Mounting, galv. iron or bronze; fits ¾″ to 1⅜″ I.D. masts.
Type 1X Base Insulator, galv. iron or bronze; top tapped standard ¾″ 16 thread.
Type 1H Base Insulator; similar to Type 1 excepting post is hinged.
Type 2 Base Insulator; galv. iron; fits ¾″ or 13/16″ I.D. masts.
Type 6 Deck Insulator; galv. iron or bronze, for deck or rooftop; fits ¾″ to 1⅜″ I.D. mast.
Type 10-S Standoff; heavy duty; chrome plated; fits ⅞″ to 1¼″ O.D. mast.
Type 8-C Insulated Mounting Clamp for horizontals or verticals; fits ⅝″ to 1″ O.D.
Type 3 Standoff Insulator for verticals or horizontals; fits ½″ to 1½″ O.D.
Type 13-S Standoff Insulator; heavy duty; fits ¾″ to 1½″ O.D.
Type 13-H Standoff Insulator, same as Type 13-S but with hinged cap.
Type 9-C Insulated Mounting Clamp for horizontals or verticals; fits ⅝″ to 1″ O.D
Type 10-C Insulated Mounting Clamp for horizontals or verticals; fits ⅝″ to 1″ O.D.
Type 7 Standoff Insulator for verticals or horizontals; fits ⅝″ to 1″ O.D.

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Inexpensive Jointed Step-Taper Whips in stai less steel and cadmium-plated oil-temper steel, in 72″, 78″, 84″, 90″ and 96″ lengt Ask for Type A.

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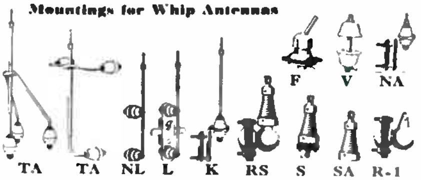
TA Mounting for trunk, panel or roof.
XL Mounting for panel of car.
L Bumper Mounting; 10″ adjustment.
K Bumper Mounting. 10″ adjustment.
RS Universal Mounting for any surface, spri style.
S Roof Mounting, spring style.
SA Spring Adaptor; use with any mounting
R-1 Universal Mounting for any surface.
F Universal Mounting; 30° adjustment
V Mounting for thru-deck or roof.
NA Bumper Mounting, 2″ adjustment
(All Fit Any ¼″ Whip Antenna)

Mountings for Whip Antennas



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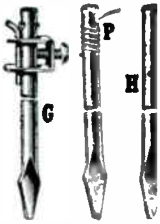

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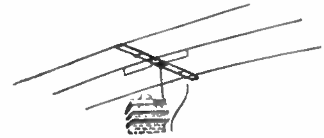

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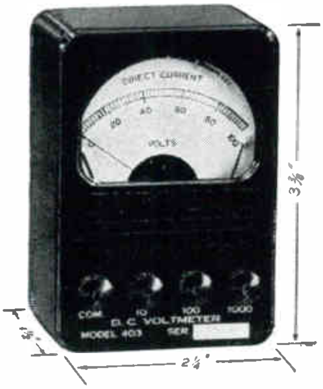

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MODEL 600 TUBE, BATTERY AND SET TESTER—Same as above with multi-meter functions...................... **$74.95**

MODEL 650 OSCILLOSCOPE—The ideal general purpose instrument for waveform observation, modulation testing and receiver alignment. Both vertical and horizontal amplifiers will pass 60 cycle square wave and have excellent response to 100,000 cycles. Sensitivity 0.3 Volts. Time base (sweep) range 20 to 30,000 cycles. All controls on front panel. Internal or External synchronization—Positive locking of pattern. 3 inch Cathode Ray tube. Size 13 x 11½ x 8½ inches. For 110–125 volts 50–60 cycle power................ **$49.95**

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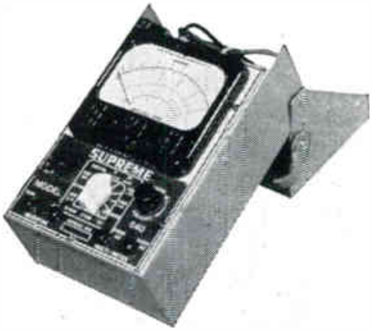
MODEL 543 VOLT-OHM-MILLIAMMETER—Popular as a companion tester among radio amateurs and technicians for general testing. Checks VOLTAGE (a-c, d-c) to 3000 volts, CURRENT (d-c) to 600 M.A., RESISTANCE to 200,000 ohms. Three inch meter indicator with SUPREME built 1000 ohms/volt movement and sturdy pointer. Available in bakelite case (Type B) for $22.45 or in metallic case (Type M) with meter protector shield at.......................... **$25.75**

MODEL 680 AUDIO GENERATOR—A high quality beat frequency type AF generator with frequency range from 15 cycles to 15,000 cycles. Ideal for testing modulation equipment. Output impedances 250/500/5000 ohms, center tapped for balanced input systems. Output voltage approx. 50 volts maximum. Output virtually flat over frequency range—within 1 db from 30 to 10,000 cycles down 2 db at 15,000 cycles. 13 x 9 x 6½ inches. Wgt. 17 lbs. For 110–125 volts 50–60 cycle power......................... **$59.95**

MODEL 574

MODEL 650

MODEL 640



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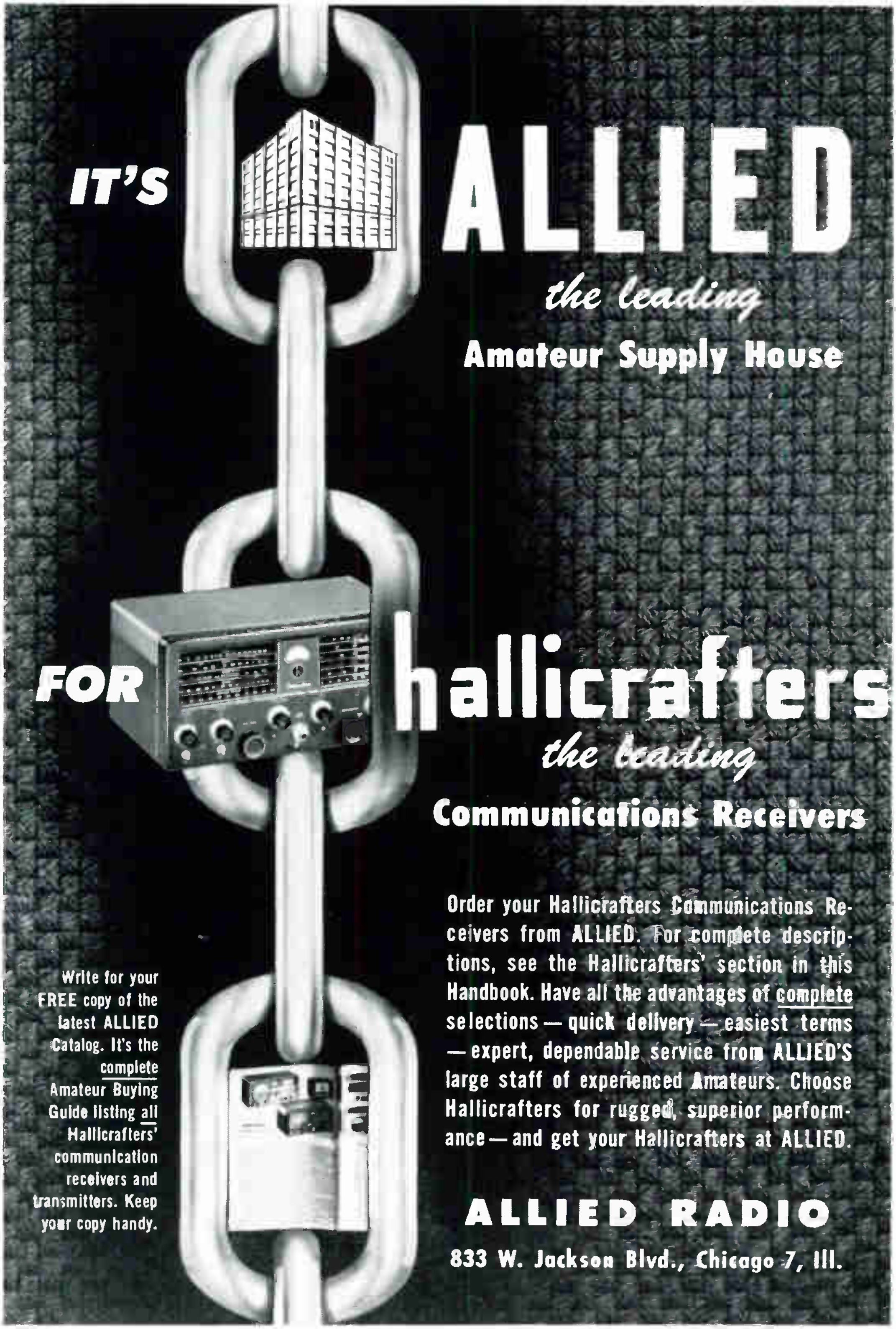

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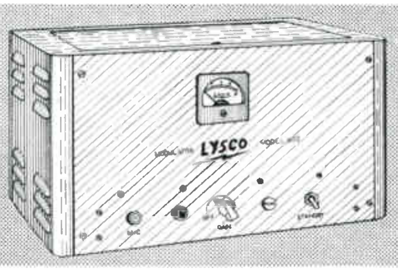

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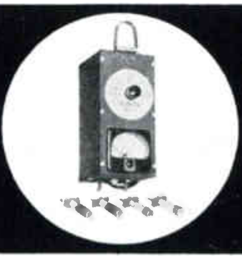

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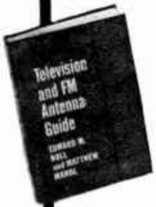

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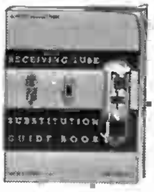

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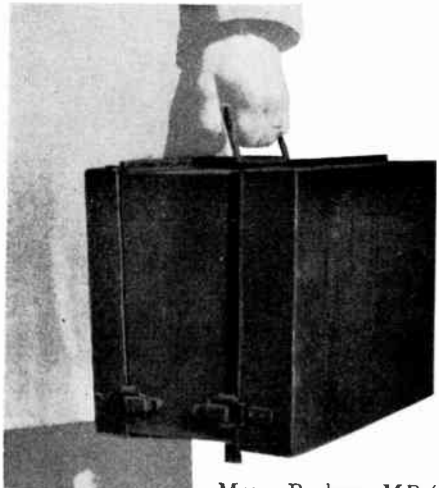

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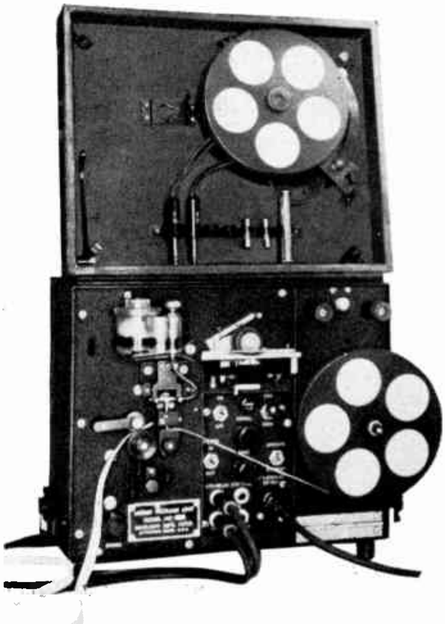

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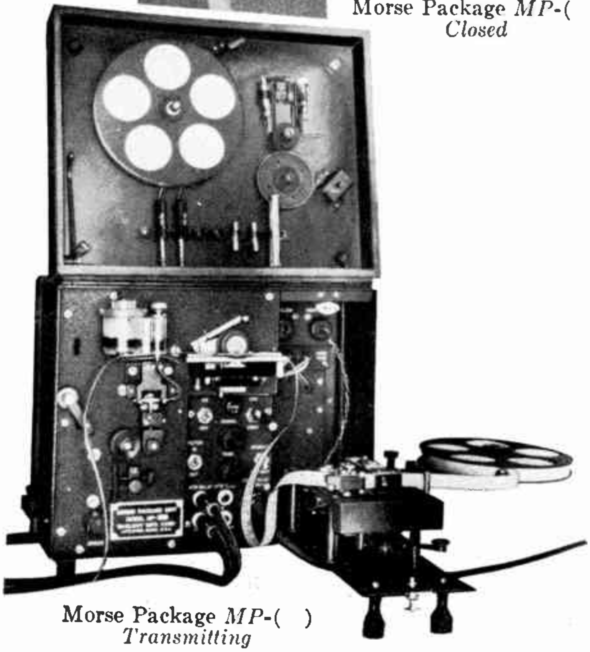

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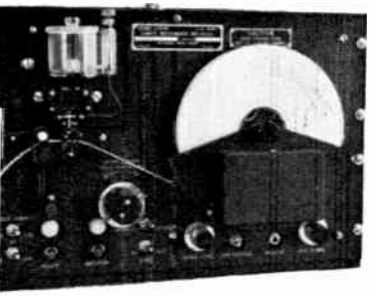

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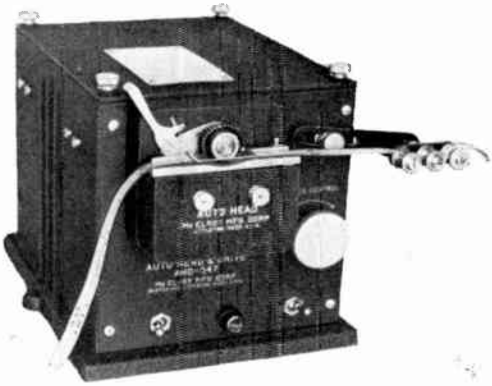

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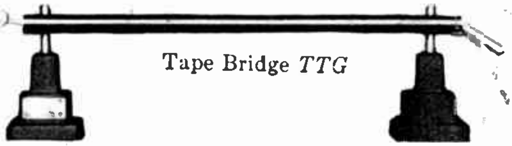

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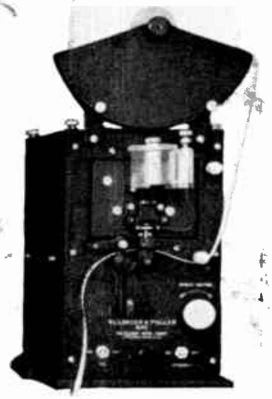

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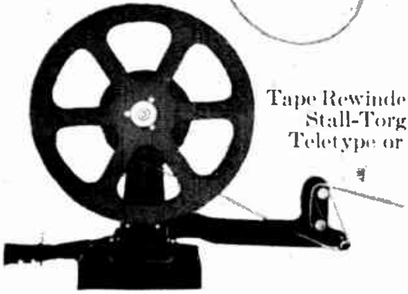

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Charts and Tables

The Radio Amateur best justifies his existence by the service he renders his community in times of disaster and distress when all other media of summoning assistance have failed. In these days of existing and pending national emergency, your services as an amateur in this respect are needed more than ever before in the increased demand for emergency communications facilities for civil defense work. The amateur's emergency service is implemented and centralized by the Amateur Radio Emergency Corps, sponsored by ARRL in the public interest. Any licensed amateur in the U. S. or Canada is invited to register his facilities in the AREC. Use the form below or facsimile thereof.

AMATEUR RADIO EMERGENCY CORPS

REGISTRATION FORM

In registering in the Amateur Radio Emergency Corps, it should be understood that there is no expense of any kind involved and that there is no obligation to join any amateur radio organization in order to retain membership

PLEASE PRINT OR TYPE

Name .... Call .... Phone ....

Home Address .... City ....

Business Address ....

Business Hours .... Days Off .... Business Phone No. ....

Have you ever handled traffic during an emergency? .... If so please describe briefly on other side. Are you a member of any military reserve organization(s) ? .... Which? ....

TRANSMITTERS

Fill in for each transmitter

Band(s) Mc.	Line-Up	Power		Port.	Mble.	Fixed	Xtal or VFO	CW or Phone
		Emerg.	Reg.					

POWER SUPPLIES

Emergency	Regular (Commercial Mains)

RECEIVERS

	Power Source (check which)	
	Emerg.	Reg.

☐ I intend to participate ACTIVELY in the AREC program and request FULL MEMBERSHIP.

☐ As my participation must be limited, as explained on reverse side, I request SUPPORTING MEMBERSHIP.

It is understood and agreed that my membership grade (FULL or SUPPORTING) will be determined by the Emergency Coordinator, Section Emergency Coordinator or Section Communications Manager supervising my area and that it will be based entirely upon the degree of my activity in local AREC plans and drills. I further understand that my continued interest and activity, together with annual endorsement of my membership card are the sole requirements for retaining my membership in good standing.

Date .... Signed ....

Forward this registration form to your EC, SEC or SCM, in that order of preference. SCM addresses are given on page 6 of each issue of *QST*. Addresses of SEC's and EC's are on file both with hq. and the SCM.