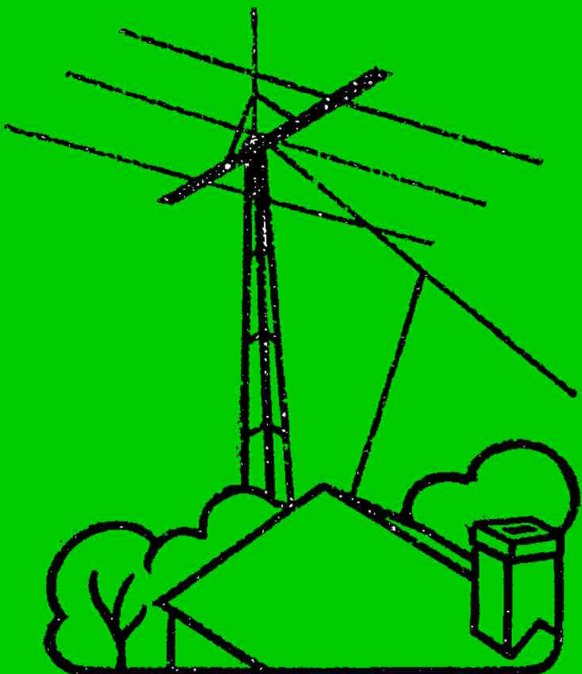
37TH EDITION · 1960

The radio amateur's handbook

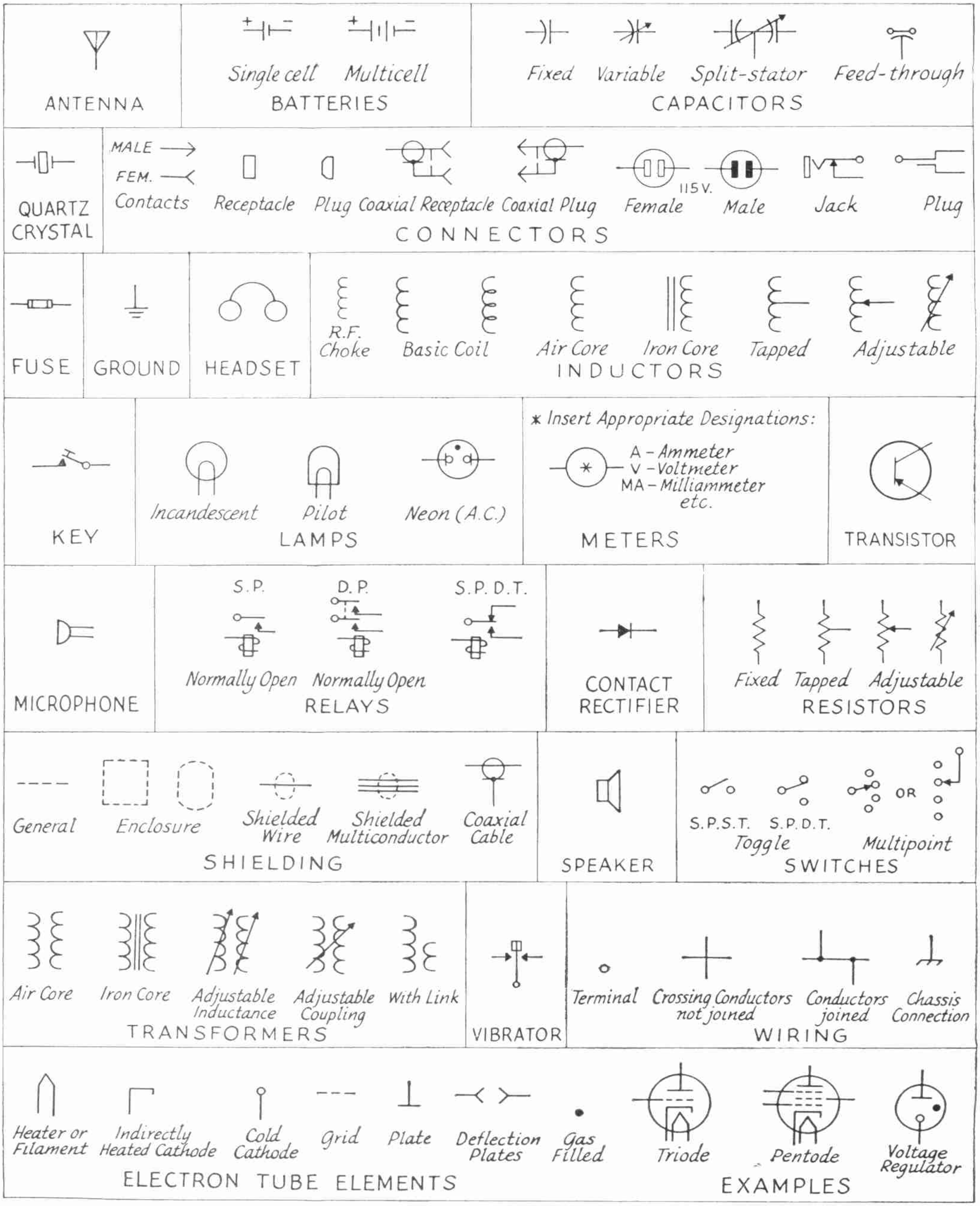
THE STANDARD MANUAL OF AMATEUR
RADIO COMMUNICATION

$3.50
U.S.A. Proper



UBLISHED BY THE AMERICAN RADIO RELAY LEAGUE

SCHEMATIC SYMBOLS USED IN CIRCUIT DIAGRAMS



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

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

In the case of switches, jacks, 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.

1960

Thirty-seventh Edition

Library of Congress Catalog Card Number: 41-3345

Thirty-seventh Edition

$3.50 in U. S. A proper
$4.00 in U. S. Possessions and Canada
$4.50 elsewhere
Buckram Edition $6.00 Everywhere

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

Foreword

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

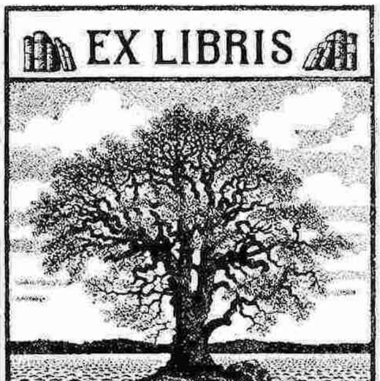
Virtually continuous modification is a feature of the *Handbook* — always with the objective of presenting the soundest and best aspects of current practice rather than the merely new and novel. Its annual revision, a major task of the headquarters group of the League, is participated in by skilled and experienced amateurs well acquainted with the practical problems in the art.

The *Handbook* is printed in the format of the League's monthly magazine, *QST*. This, together with extensive and useful catalog advertising by manufacturers producing equipment for the radio amateur and industry, 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.

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

A. L. Budlong
General Manager, A.R.R.L.

West Hartford, Conn.


EX LIBRIS
RJ

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 cooperation 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*

Amateur Radio

Amateur radio is a scientific hobby, a means of gaining personal skill in the fascinating art of electronics and an opportunity to communicate with fellow citizens by private short-wave radio. Scattered over the globe are over 250,000 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 over 200,000, 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 military services seek the cooperation 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; traditional amateur skills in emergency communication are also the stand-by system for the nation's civil defense. 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 homemade spark transmitters. By 1912 there were numerous Government and commercial stations, and hundreds of amateurs; regulation was needed, so laws, licenses and wavelength specifications 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 1000-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 War I not only marked the close of the first phase of amateur development but came very

HIRAM PERCY MAXIM
President ARRL, 1914–1936

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

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

TRANSATLANTICS

As DX became 1000, then 1500 and then 2000 miles, amateurs began to dream of transatlantic 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 transatlantic 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*. Excitement began to spread through amateur ranks.

Finally, in November, 1923, after some months of careful preparation, two-way amateur transatlantic communication was accomplished, when Schnell, 1MO, and Reinartz, 1XAM (now W4CF 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, 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.

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 military and civil defense authorities 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 Affiliate 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 so outstanding that other explorers followed suit. During subsequent years a total of perhaps two hundred voyages and expeditions were assisted by amateur radio, the several explorations of the Antarctic being perhaps the best known.

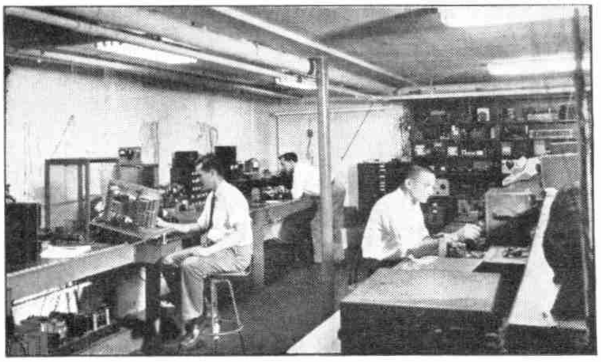
Since 1913 amateur radio has been the principal, and in many cases the only, means of outside communication in several hundred storm, flood and earthquake emergencies in this country. The 1936 and 1937 eastern states floods, the Southern California flood and Long Island-New England hurricane disaster in 1938, the Florida-Gulf Coast hurricanes of 1947, and the 1955 flood disasters 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 cooperation with the Red Cross, and in 1947 a National Emergency Coordinator was appointed to full-time duty at League headquarters.

The amateur's outstanding record of organized preparation for emergency communications and performance under fire has been largely responsible for the decision of the Federal Government to set up special regulations and set aside special frequencies for use by amateurs in providing auxiliary communications for civil defense purposes in the event of war. Under the banner, "Radio Amateur Civil Emergency Service," amateurs are setting up and manning community and area networks integrated with civil defense functions of the municipal governments. Should a war cause the shut-down of routine amateur activities, the RACES will be immediately available in the national defense, manned by amateurs highly skilled in emergency communication.

● 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 not uncommon; during solar peaks, 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 "single-sideband suppressed-carrier" systems as well as even more selectivity in receiving equipment for greater efficiency in spectrum use.

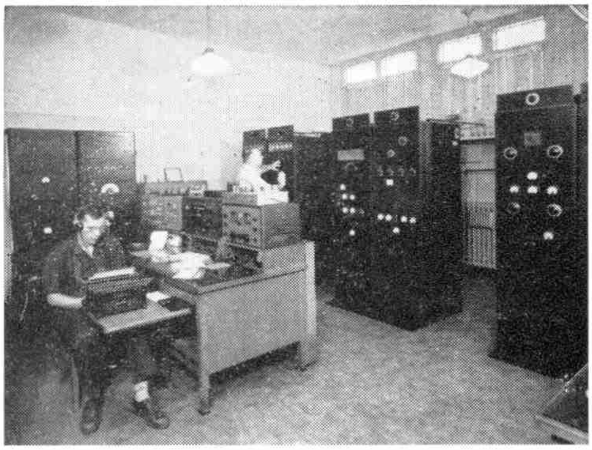
During World War II, thousands of skilled amateurs contributed their knowledge to the development of secret radio devices, both in Government and private laboratories. Equally as important, the prewar technical progress by amateurs provided the keystone for the development of modern military communications equipment. Perhaps more important today than individual contributions to the art is the mass cooperation of the amateur body in Government projects such as propagation studies; each participating station is in reality a separate field laboratory from which reports are made for correlation and analysis. An outstanding example was varied amateur participation in several activities of the International Geophysical Year program. ARRL, with Air Force sponsorship, conducted an intensive study of v.h.f. propagation phenomena — DX transmissions via little-understood methods such as meteor and auroral reflections, and transequatorial scatter. ARRL-affiliated clubs and groups have operated precision receiving antennas and apparatus to help track earth satellites via radio. For volunteer astronomers searching visually for the satellites, other amateurs have manned networks to provide instant radio reports of sightings to a central agency so that an orbit might be computed.

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.



The operating room at W1AW.

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 cooperative 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 territory of ARRL is divided into one Canadian and fifteen U. S. 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 one by the Canadian membership. These directors then choose the president and vice-president, who are also members of the Board. The secretary and treasurer are also 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. The directors appoint a general manager to supervise the operations of the League and its headquarters, and to carry out the policies and instructions of the Board.

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-three sections. There are appointments for qualified members in various fields, as outlined in Chapter 24. 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

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. There are four available classes of amateur license — Novice, Technician, General (called "Conditional" if exam taken by mail), and Amateur Extra Class. Each has different requirements, the first two being the simplest and consequently conveying limited privileges as to frequencies available. Exams for Novice, Technician and Conditional classes are taken by mail under the supervision of a volunteer examiner. 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; some are available for radiotelephone, others for special forms of transmission such as teletype, facsimile, amateur television or radio control. 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 subject to further regulations. 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 code at the required speed, 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 by an FCC engineer (or by a volunteer, depending on the license class), through FCC at Washington. A complete up-to-the-minute discussion of license requirements, and study guides for those preparing for the examinations, 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 50¢, postpaid.

LEARNING THE CODE

In starting to learn the code, you should consider it simply another means of conveying

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: dahdididi-dah. Wait: didahdididit. End of message: didahdidahdit. Invitation to transmit: dahdidah. End of work: didididahdidah. Fraction bar: dahdididahdit.

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

information. The spoken word is one method, 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. Learning the code is not at all hard; a simple booklet treating the subject in detail is another of the beginner publications available from the League, and is entitled, *Learning the Radiotelegraph Code*, 50¢ postpaid.

Code-practice transmissions are sent by W1AW every evening at 2130 EST (EDST May through October). See Chapter 24, "Code Proficiency."

THE AMATEUR BANDS

Amateurs are assigned bands of frequencies at approximate harmonic intervals throughout the spectrum. Like assignments to all services, they are subject to modification to fit the changing picture of world communications needs. Modifications of rules to provide for domestic needs are also occasionally issued by FCC, and in that respect each amateur should keep himself informed by W1AW bulletins, *QST* reports, or by communication with ARRL Hq. concerning a specific point.

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 tone-modulated c.w. telegraphy, A3 is amplitude-modulated phone, A4 is facsimile, A5 is television, n.f.m. designates narrow-band frequency- or phase-modulated radiotelephony, f.m. means frequency modulation, phone (including n.f.m.) or telegraphy, and F1 is frequency-shift keying.

80 meters	3.500–4.000	— A1
	3.500–3.800	— F1
	3.800–4.000	— A3 and n.f.m.
40 m.	7.000–7.300	— A1
	7.000–7.200	— F1
	7.200–7.300	— A3 and n.f.m.
20 m	14.000–14.350	— A1
	14.000–14.200	— F1
	14.200–14.300	— A3 and n.f.m.
	14.300–14.350	— F1
15 m.	21.000–21.450	— A1
	21.000–21.250	— F1
	21.250–21.450	— A3 and n.f.m.
10 m.	28.000–29.700	— A1
	28.500–29.700	— A3 and n.f.m.
	29.000–29.700	— f.m.
6 m.	50–54	— A1, A2, A3, A4, n.f.m.
	51–54	— A0
	52.5–54	— f.m.
2 m.	144–148 220–225	— A1, A0, A1, A2, A3, A4, f.m.
	420–450[1] 1,215–1,300	A0, A1, A2, A3, A4, A5, f.m.
	2,300– 2,450 3,500– 3,700 5,650– 5,925 10,000–10,500[2] 21,000–22,000 All above 30,000	A0, A1, A2, A3, A4, A5, f.m., pulse

[1] Input power must not exceed 50 watts.

[2] No pulse permitted in this band.

NOTE: The bands 220 through 10,500 Mc. are shared with the Government Radio Positioning Service, which has priority.

In addition, A1 and A3 on portions of 1.800–2.000, as follows:

		Power (watts)	
Area	*Band ,kc.*	*Day*	*Night*
Minn., Iowa, Wis., Mich., Pa., Md., Del. and states to north	1800–1825	500	200
N.D., S.D., Nebr., Colo., N. Mex., and states west, including Hawaiian Ids.	1975–2000	500*	200*
Okla., Kans., Mo., Ark., Ill., Ind., Ky., Tenn., Ohio, W. Va., Va., N. C., S. C., and Texas (west of 99° W *or* north of 32° N)	1800–1825	200	50

No operation elsewhere.

* Except in state of Washington, 200 watts day, 50 watts night.

Novice licensees may use the following frequencies, transmitters to be crystal-controlled and have a maximum power input of 75 watts.

3.700–3.750	A1	21.100–21.250	A1
7.150–7.200	A1	145–147	A1, A2, A3, f.m.

Technician licensees are permitted all amateur privileges in 50-54 Mc., 145-147 Mc., and in the bands 220 Mc. and above.

CHAPTER 2

Electrical Laws and Circuits

ELECTRIC AND MAGNETIC FIELDS

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**. In radio work, 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. The field exerts a *force* on an object immersed in it; this force represents potential (ready-to-be-used) energy, so the **potential** of the field is a measure of the **field intensity.** The **direction** of the field is 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 neither moves nor changes in intensity. 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

Although no one knows what it is that composes the field itself, it is useful to invent a picture of it that will help in visualizing the forces and the way in which they act.

A field can be pictured as being 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 unit of area (square inch or square centimeter) is called the **flux density.**

ELECTRICITY AND THE ELECTRIC CURRENT

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 several different kinds of still smaller particles. One is the **electron,** essentially a small particle of electricity. The quantity or **charge** of electricity represented by the electron is, in fact, the smallest quantity of electricity that can exist. The kind of electricity associated with the electron is called **negative.**

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. The nucleus has an electric charge of the kind of electricity called **positive,** the amount of its charge being just exactly equal to the sum of the negative charges on all the electrons associated with that nucleus.

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 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.

In a normal atom the positive charge on the nucleus is exactly balanced by the negative charges on the electrons. However, it is possible for an atom to lose one of its electrons. When that happens the atom has a little less negative charge than it should — that is, 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 possible for electrons to travel from atom to atom. The movement of ions or electrons constitutes the **electric current.**

The **amplitude** of the current (its intensity or magnitude) is determined by the rate at which electric charge — an accumulation of electrons or ions of the same kind — moves past a point in a circuit. Since the charge on a single electron or ion is extremely small, the number that must move as a group to form even a tiny current is almost inconceivably large.

Conductors and Insulators

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 list 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 or potential (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.

In picturing current flow it is natural to think of a single, constant force causing the electrons to move. When this is so, the electrons 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 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 to begin the next cycle. 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 strength 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

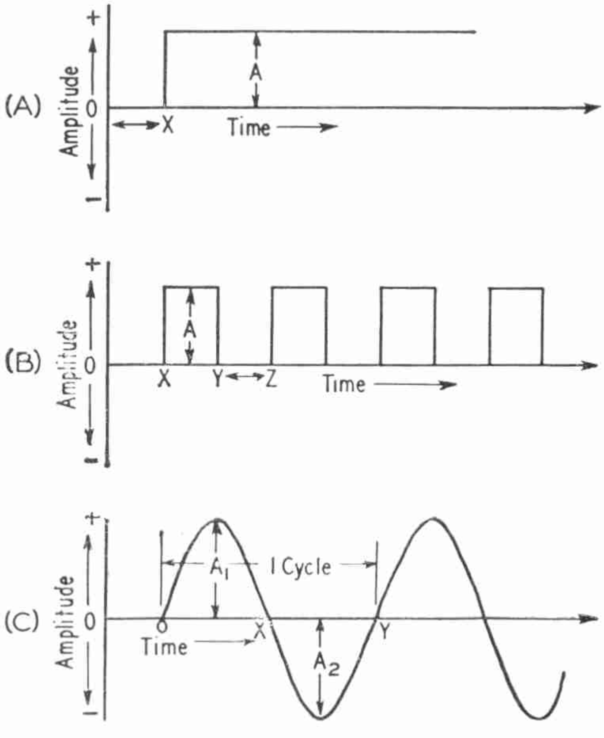


Fig. 2-1—Three types of current flow. A—direct current; B—intermittent direct current; C—alternating current.

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.

Waveforms

The type of alternating current shown in Fig. 2-1C is known as a **sine wave.** 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 **complex waves** 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. 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 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 equal one ampere.

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 sine-wave a.c., this **effective** (or **r.m.s.**) **value** is equal to the *maximum* amplitude (A_1 or A_2 in Fig. 2-1C) multiplied by 0.707. The **instantaneous value** is the value that the current (or voltage) has at any selected instant in the cycle.

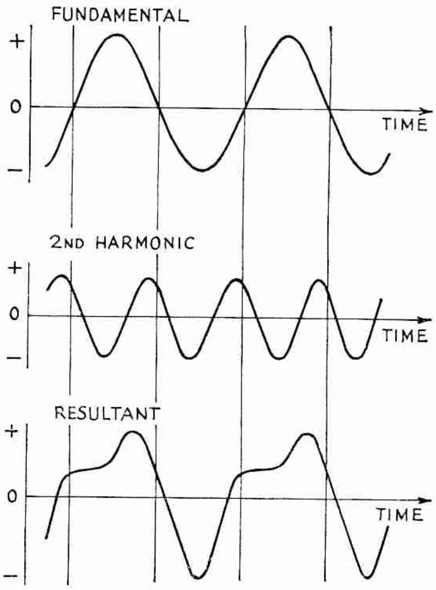


Fig. 2-2—A complex waveform. A fundamental (top) and second harmonic (center) added together, point by point at each instant, result in the waveform shown at the bottom. When the two components have the same polarity at a selected instant, the resultant is the simple sum of the two. When they have opposite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

If all the instantaneous values in a sine wave are averaged over a *half*-cycle, the resulting figure is the **average value.** 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.

● FREQUENCY AND WAVELENGTH

Frequency Spectrum

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 a similar 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 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 abbreciated **Mc.**

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

Radio waves travel at the same speed as light — 300,000,000 meters or about 186,000 miles a second in space. They can be 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 an r.f. current has a frequency of 3,000,000 cycles per second. The fields 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. 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 **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

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

Resistance

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 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, and 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 various conductors 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.

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 of Wires

The problem of determining 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 — can be easily solved with the help of the copper-wire table given in a later 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 Chapter 20 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

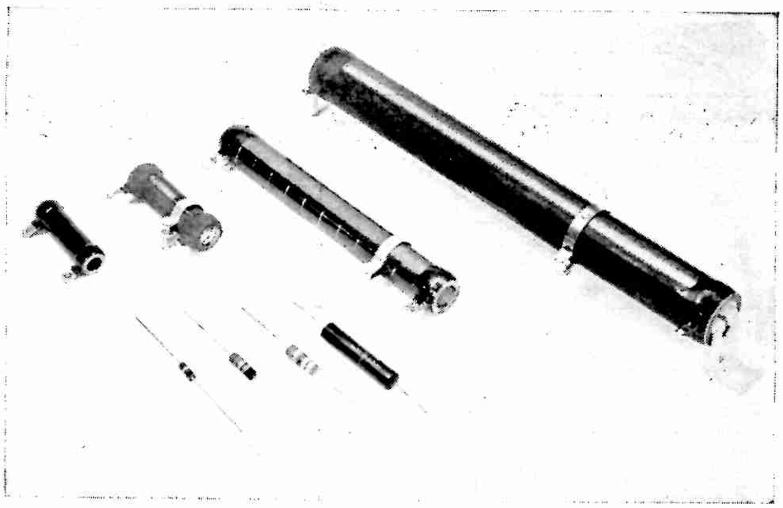
$$\frac{14}{1000} \times R = 0.05 \text{ ohm}$$

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 multi-



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 of the adjustable type, having a sliding contact on an exposed section of the resistance winding.

plied 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 resistance calculations for 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

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. 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 the heat quickly it may reach a temperature that will cause it to melt or burn.

Skin Effect

The resistance of a conductor is not the same for alternating current as it is for direct current. When the current is alternating there are internal effects that tend to force the current to flow mostly in the outer parts of the conductor. This decreases the effective cross-sectional area of the conductor, with the result that the resistance increases.

For low audio frequencies the increase in resistance is unimportant, but at radio frequencies this **skin effect** is so great that practically all the current flow is confined within a few thousandths of an inch of the conductor surface. The r.f. resistance is consequently many times the d.c. resistance, and increases with increasing frequency. In the r.f. range a conductor of thin tubing will have just as low resistance as a solid conductor of the same diameter, because material not close to the surface carries practically no current.

Conductance

The reciprocal of resistance (that is, $1/R$) is called **conductance.** It is usually represented by the symbol G. 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.

● 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.

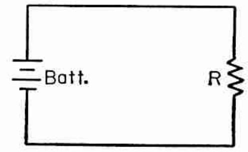


Fig. 2-3—A simple circuit consisting of a battery and resistor.

TABLE 2-II

Conversion Factors 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

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 each 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 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.

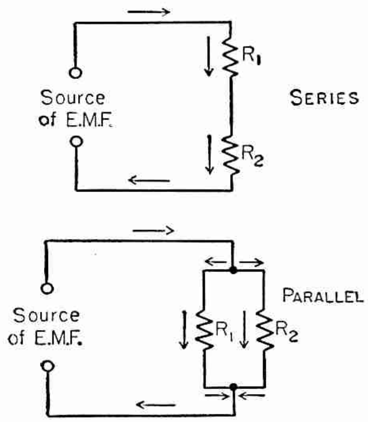


Fig. 2-4—Resistors connected in series and in parallel.

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.**

Series and Parallel Resistance

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 among them. The voltage appearing across each resistor (the **voltage drop**) 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

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

The applied 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 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.

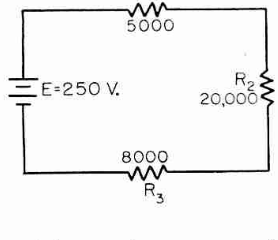


Fig. 2-5—An example of resistors in series. The solution of the circuit is worked out in the text.

Resistors in Parallel

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

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

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 becomes

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

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

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

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

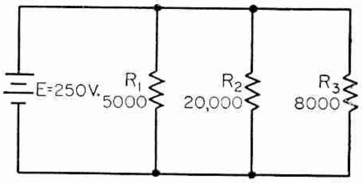


Fig. 2-6—An example of resistors in parallel. The solution is worked out in the text.

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

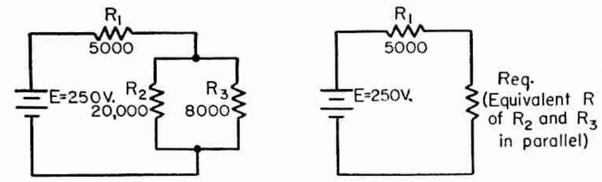


Fig. 2-7—An example of resistors in series-parallel. The equivalent circuit is at the right. The solution is worked out in the text.

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

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

The total resistance in the circuit is then

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

The current is

$$I = \frac{E}{R} = \frac{250}{10.71} = 23.3 \text{ ma.}$$

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

$$E_1 = IR_1 = 23.3 \times 5 = 117 \text{ volts}$$
$$E_2 = IR_{\text{eq.}} = 23.3 \times 5.71 = 133 \text{ volts}$$

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

$$I_2 = \frac{E_2}{R_2} = \frac{133}{20} = 6.65 \text{ ma.}$$
$$I_3 = \frac{E_2}{R_3} = \frac{133}{8} = 16.6 \text{ ma.}$$

where I_2 = Current through R_2
I_3 = Current through R_3

The total is 23.25 ma., which checks closely enough with 23.3 ma., the current through the whole circuit.

● 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:

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

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

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

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. 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.

Generalized Definition of Resistance

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. But in every case of this kind the power is completely "used up" — it cannot be recovered. Also, for proper operation of the device the power must be supplied at a definite ratio of voltage to current. Both these features are characteristics of resistance, so it can be said that any device that dissipates power has a definite value of "resistance." This concept of resistance as something that absorbs power at a definite voltage/current ratio is very useful, since it permits substituting a simple resistance for the **load** or power-consuming part of the device receiving power, often with considerable simplification of calculations. Of course, every electrical device has some resistance of its own in the more narrow sense, 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.

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

Suppose two flat metal plates are placed close to each other (but not touching) and are connected to a battery through a switch, as shown in Fig. 2-8. 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. 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.

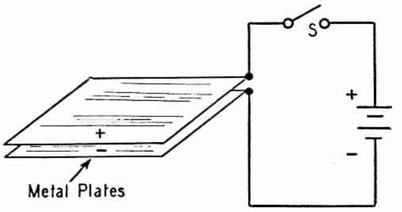


Fig. 2-8—A simple capacitor.

If the switch is opened after the plates have been **charged** in this way, the top plate is left with a deficiency of electrons and the bottom plate with an excess. The plates remain charged despite the fact that the battery no longer is connected. 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. The plates have then been **discharged.**

The two plates constitute an electrical **capacitor,** and from the discussion above it should be clear that a capacitor possesses the property of storing electricity. (The energy actually is stored in the electric field between the plates.) It should also be clear that during the time the electrons are moving — that is, while the capacitor is being charged or discharged — a current is flowing in the circuit even though the circuit is "broken" by the gap between the capacitor 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 capacitor.

The **charge** or quantity of electricity that can be placed on a capacitor is proportional to the applied voltage and to the **capacitance** of the capacitor. The larger the plate area and the smaller the spacing between the plate the 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 many times. The ratio of the capacitance with some material other than air between the plates, to the capacitance of the same capacitor 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 capacitors are given in Table 2-III. If a sheet of photographic glass is substituted for air between the plates of a capacitor, for example, the capacitance will be increased 7.5 times.

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
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
Teflon	1.9–2.6	700–1100
Wood (dry oak)	2.5–6.8	

* In volts per mil (0.001 inch).

Unit

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 μf.) or **micromicrofarads** (μμf.). The microfarad is one-millionth of a farad, and the micromicrofarad is one-millionth of a microfarad. Capacitors 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, 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.

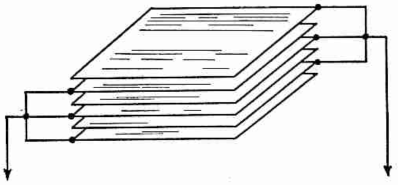


Fig. 2-9—A multiple-plate capacitor. Alternate plates are connected together.

The formula for calculating capacitance is:

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

where C = Capacitance in μμf.
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.

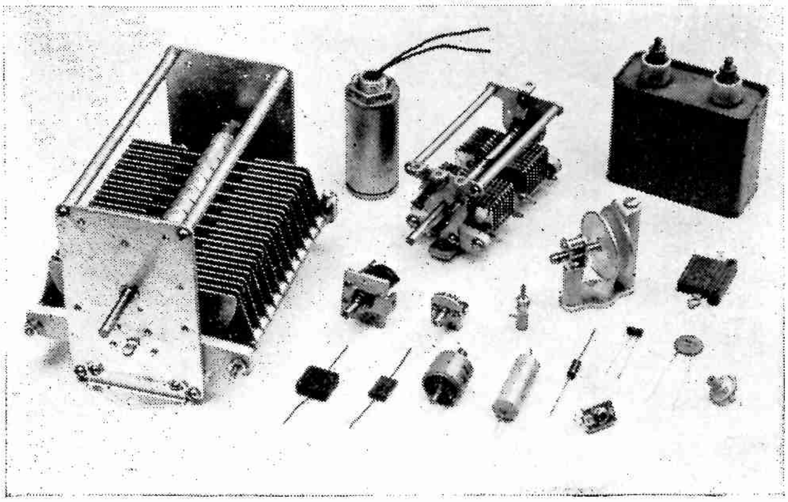
The usefulness of a capacitor in electrical circuits lies in the fact that it can be charged with electrical energy at one time and then discharged at a later time. In other words, it is an "electrical reservoir."

Capacitors in Radio

The types of capacitors used in radio work differ considerably in physical size, construction, and capacitance. Some representative types are shown in the photograph. In **variable** capacitors (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** capacitors — that is, assemblies having a single, non-adjustable value of 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, paper and special ceramics. An example of a liquid dielectric is mineral oil. The **electrolytic** capacitor 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 capacitor. The capacitance obtained with a given plate area in an electrolytic capacitor is very large, compared with capacitors 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 capacitor, 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 Table 2-III. 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



Fixed and variable capacitors. The large unit at the left is a transmitting-type variable capacitor for r.f. tank circuits. To its right are other air-dielectric variables of different sizes ranging from the midget "air padder" to the medium-power tank capacitor at the top center. The cased capacitors in the top row are for power-supply filters, the cylindrical-can unit being an electrolytic and the rectangular one a paper-dielectric capacitor. Various types of mica, ceramic, and paper-dielectric capacitors are in the foreground.

evidenced by a spark or arc between the plates, but if the voltage is removed the arc ceases and the capacitor 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 capacitor must have more plate area than a low-voltage one of the same capacitance. High-voltage high-capacitance capacitors are physically large.

CAPACITORS IN SERIES AND PARALLEL

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

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

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

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

and, for only two capacitors in series,

$$C\ (\text{total}) = \frac{C_1C_2}{C_1 + C_2}$$

The same units must be used throughout; that is, all capacitances must be expressed in either μf. or $\mu\mu$f.; both kinds of units cannot be used in the same equation.

Capacitors 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 capacitors in parallel is the voltage that can be applied safely to the one having the *lowest* voltage rating.

When capacitors are connected in series, the applied voltage is divided up among them; 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 capacitor of a group connected in series is in *inverse* proportion to its capacitance, as compared with the capacitance of the whole group.

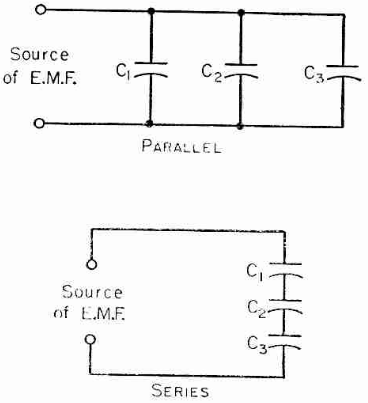


Fig. 2-10—Capacitors in series and parallel.

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

$$C = \cfrac{1}{\cfrac{1}{C_1} + \cfrac{1}{C_2} + \cfrac{1}{C_3}} = \cfrac{1}{\cfrac{1}{1} + \cfrac{1}{2} + \cfrac{1}{4}} = \cfrac{1}{\cfrac{7}{4}} = \frac{4}{7}$$
$$= 0.571\ \mu\text{f}.$$

The voltage across each capacitor is proportional to the *total* capacitance divided by the capacitance of the capacitor 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.

Capacitors are frequently connected in series to enable the group to withstand a larger voltage (at the expense of decreased total capacitance) than any individual capacitor is rated to stand. However, as shown by the previous example, the applied voltage does not divide equally among the capacitors (except when all the capacitances are the same) so care must be taken to see that the voltage rating of no capacitor in the group is exceeded.

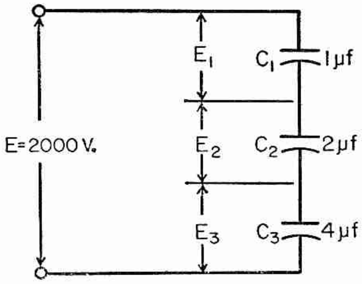


Fig. 2-11—An example of capacitors connected in series. The solution to this arrangement is worked out in the text.

2—ELECTRICAL LAWS AND CIRCUITS

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 current, in other words, sets up a magnetic field.

The transfer of energy to the magnetic field represents work done by the source of e.m.f. Power is required for doing work, and since power is equal to current multiplied by voltage, there must be a voltage drop in the circuit during the time in which energy is being stored in the field. This voltage "drop" (which has nothing to do with the voltage drop in any resistance in the circuit) is the result of an opposing voltage "induced" in the circuit while the field is building up to its final value. When the field becomes constant the **induced e.m.f.** or **back e.m.f.** disappears, since no further energy is being stored.

Since the induced e.m.f. opposes the e.m.f. of the source, it tends to prevent the current from rising rapidly when the circuit is closed. The amplitude of the induced e.m.f. is proportional to the rate at which the current is changing and to a constant associated with the circuit itself, called the **inductance** of the circuit.

Inductance depends on the physical characteristics of the conductor. If the conductor is formed into a coil, for example, its inductance is increased. A coil of many turns will have more inductance than one of few turns, if both coils are otherwise physically similar. Also, if a coil is placed on an iron core its inductance will be greater than it was without the magnetic core.

The polarity of an induced e.m.f. is always such as to oppose any change in the current in the circuit. This means that when the current in the circuit is increasing, work is being done against the induced e.m.f. by storing energy in the magnetic field. If the current in the circuit tends to decrease, the stored energy of the field returns to the circuit, and thus adds to the energy being supplied by the source of e.m.f. This tends to keep the current flowing even though the applied e.m.f. may be decreasing or be removed entirely.

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 of the "air-core" type; that is, wound on an insulating support consisting of nonmagnetic material.

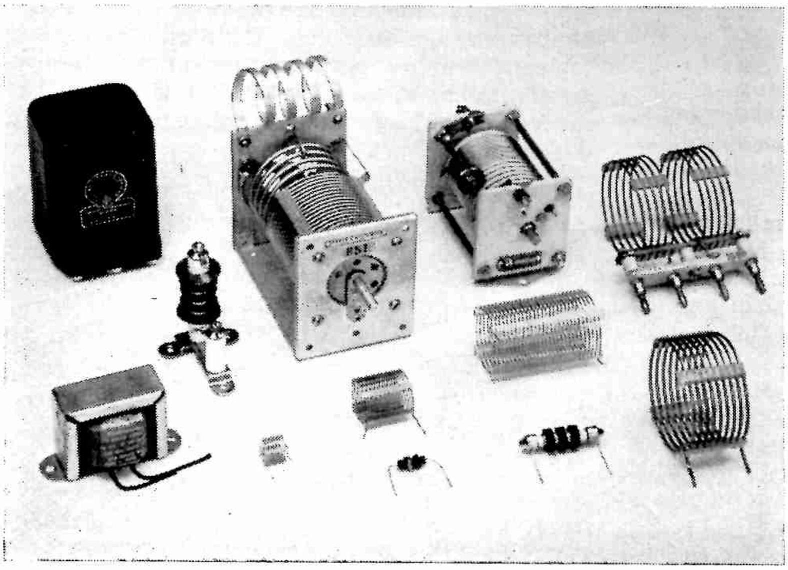
Every conductor has inductance, even though the conductor is not formed into a coil. The inductance of a short length of straight wire is small, but it may not be negligible because if the current through it changes its intensity rapidly enough the induced voltage may be appreciable. This will be the case in even a few inches of wire when an alternating current having a frequency of the order of 100 Mc. or higher 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.

Calculating Inductance

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



Inductors for power and radio frequencies. The two iron-core coils at the left are "chokes" for power-supply filters. The mounted air-core coils at the top center are adjustable inductors for transmitting tank circuits. The "pie-wound" coils at the left and in the foreground are radio-frequency choke coils. The remaining coils are typical of inductors used in r.f. tuned circuits, the larger sizes being used principally for transmitters.

c = Radial depth of winding in inches
n = Number of turns

The notation is explained in Fig 2-12. The quantity $10c$ may be neglected if the coil only has one layer of wire.

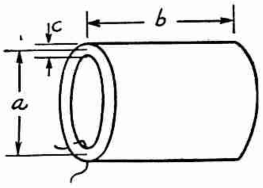


Fig. 2-12 — Coil dimensions used in the inductance formula.

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, 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:

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

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

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

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

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

A 27-turn coil would be close enough to the required value of inductance, in practical work. Since the coil will be 1.25 inches long, the number of turns per inch will be 27/1.25 = 21.6. Consulting the wire table, we find that No. 18 enameled wire (or any smaller size) can be used. The proper inductance is obtained 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.

Inductance Charts

Most inductance formulas lose accuracy when applied to small coils (such as are 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. Fig. 2-13 shows the measured inductance of v.h.f. coils, 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.

The charts of Figs. 2-14 and 2-15 are useful for rapid determination of the inductance of coils of the type commonly used in radio-frequency circuits in the range 3–30 Mc. They are based on the formula above, and are of sufficient accuracy for most practical work. Given the coil length in inches, the curves show the multiplying factor to be applied to the inductance value given in the table below the curve for a coil of the same diameter and number of turns per inch.

Example: A coil 1 inch in diameter is 1¼ inches long and has 20 turns. Therefore it has 16 turns per inch, and from the table under Fig. 2-15 it is found that the reference inductance for a coil of this diameter and number of turns per inch is 16.8 μh. From curve B in the figure the multiplying factor is 0.35, so the inductance is

$$16.8 \times 0.35 = 5.9\ \mu\text{h.}$$

The charts also can be used for finding suitable dimensions for a coil having a required value of inductance.

Example: A coil having an inductance of 12 μh. is required. It is to be wound on a form having a diameter of 1 inch, the length available for the winding being not more than 1¼ inches. From Fig. 2-15, the multiplying factor for a 1-inch diameter coil (curve B) having the maximum possible length of 1¼ inches is 0.35. Hence the number of turns per inch must be chosen for a reference inductance of at least 12/0.35, or 34 μh. From the Table under Fig. 2-15 it is seen that 16 turns per inch (reference inductance 16.8 μh.) is too small. Using 32 turns per inch, the multiplying factor is 12/68, or 0.177, and from curve B this corresponds to a coil length of ¾ inch. There will be 24 turns in this length, since the winding "pitch" is 32 turns per inch.

IRON-CORE COILS

Permeability

Suppose that the coil in Fig. 2-16 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

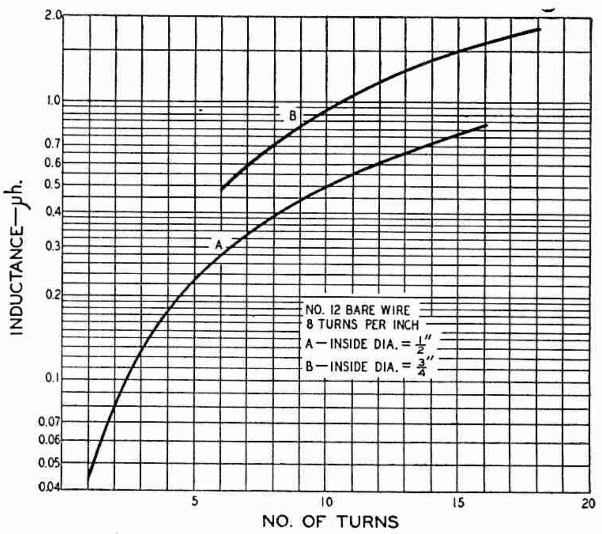


Fig. 2-13—Measured inductance of coils wound with No. 12 bare wire, 8 turns to the inch. The values include half-inch leads.

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 since, other things being equal, the inductance will be proportional to the magnetic flux through the coil.

The permeability of a magnetic material 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, but at very high flux densities, increasing the current may cause no appreciable change in the flux. 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 inductor 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.

Iron core coils such as the one sketched in Fig. 2-16 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

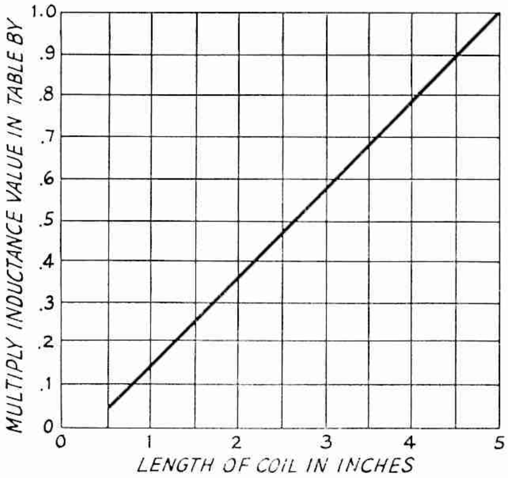


Fig. 2-14—Factor to be applied to the inductance of coils listed in the table below, for coil lengths up to 5 inches.

Coil diameter, Inches	*No. of turns per inch*	*Inductance in μh.*
1¼	4	2.75
	6	6.3
	8	11.2
	10	17.5
	16	42.5
1½	4	3.9
	6	8.8
	8	15.6
	10	24.5
	16	63
1¾	4	5.2
	6	11.8
	8	21
	10	33
	16	85
2	4	6.6
	6	15
	8	26.5
	10	42
	16	108
2½	4	10.2
	6	23
	8	41
	10	64
3	4	14
	6	31.5
	8	56
	10	89

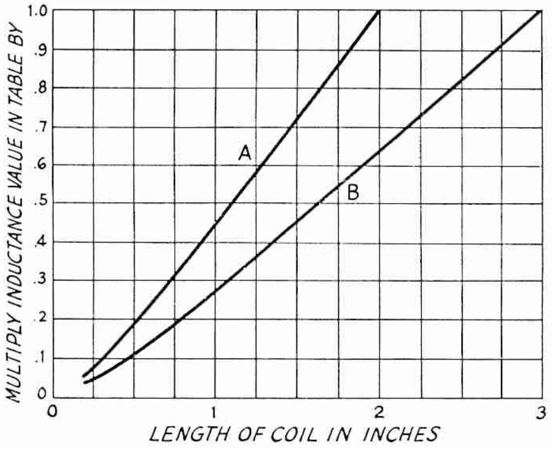


Fig. 2-15—Factor to be applied to the inductance of coils listed in the table below, as a function of coil length. Use curve A for coils marked *A*, curve B for coils marked *B*.

Coil diameter, Inches	*No. of turns per inch*	*Inductance in μh.*
½ (A)	4	0.18
	6	0.40
	8	0.72
	10	1.12
	16	2.9
	32	12
⅝ (A)	4	0.28
	6	0.62
	8	1.1
	10	1.7
	16	4.4
	32	18
¾ (B)	4	0.6
	6	1.35
	8	2.4
	10	3.8
	16	9.9
	32	40
1 (B)	4	1.0
	6	2.3
	8	4.2
	10	6.6
	16	16.8
	32	68

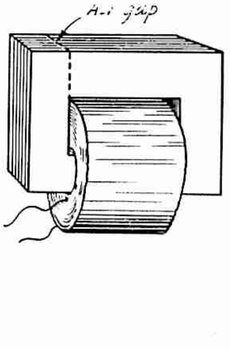


Fig. 2-16—Typical construction of an iron-core inductor. The small air gap prevents magnetic saturation of the iron and thus maintains the inductance at high currents.

the saturation point of the iron. This is done by opening 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 reduces the inductance, but makes it practically constant regardless of the value of the current.

Eddy Currents and Hysteresis

When alternating current flows through a coil wound on an iron core an e.m.f. will 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: 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, ordinary iron cores can be used only at power and audio frequencies — up to, say, 15,000 cycles. Even so, a very good grade or 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 satisfactorily 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 with 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 air, 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 inductors are connected in series (Fig. 2-17, 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,

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

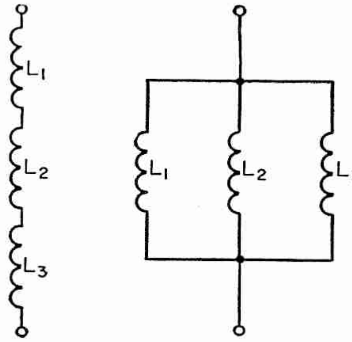


Fig. 2-17—Inductances in series and parallel.

If inductors are connected in parallel (Fig. 2-17, right), the total inductance is

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

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.

MUTUAL INDUCTANCE

If two coils are arranged with their axes on the same line, as shown in Fig. 2-18, 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.

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 theoretically be obtained with two given coils is called the **coefficient of coupling** between the coils. It is frequently expressed as a percentage. Coils that have nearly the maximum possible (coefficient = 1 or 100%) 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

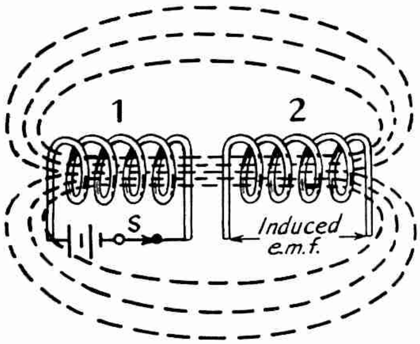


Fig. 2-18—Mutual inductance. When the switch, S, is closed current flows through coil No. 1, setting up a magnetic field that induces an e.m.f. in the turns of coil No. 2.

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 and are as close together as possible (one wound over the other). 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 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.

Time Constant

Capacitance and Resistance

Connecting a source of e.m.f. to a capacitor causes the capacitor to become charged to the full e.m.f. practically instantaneously, if there is no resistance in the circuit. However, if the circuit contains resistance, as in Fig. 2-19A, the resistance limits the current flow and an appreciable length of time is required for the e.m.f. between the capacitor plates to build up to the same value as the e.m.f. of the source. During this "building-up" period the current gradually decreases from its initial value, because the increasing e.m.f. stored on the capacitor offers increasing opposition to the steady e.m.f. of the source.

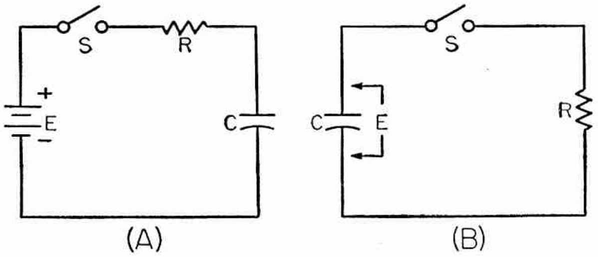


Fig. 2-19—Illustrating the time constant of an *RC* circuit.

Theoretically, the charging process is never really finished, but eventually the charging current 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 capacitor to reach 63 per cent of the applied e.m.f. (this figure is chosen for mathematical reasons). The voltage across the capacitor rises with time as shown by Fig. 2-20.

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. These units usually are more convenient.

Example: The time constant of a 2-μf. capacitor and a 250,000-ohm (0.25 megohm) resistor is

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

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

If a charged capacitor is *discharged* through a resistor, as indicated in Fig. 2-19B, the same time constant applies. If there were no resistance, the capacitor would discharge instantly when S was closed. However, since R limits the current flow the capacitor voltage cannot instantly go to zero, but it will decrease just as rapidly as the capacitor can rid itself of its charge through R. When the capacitor 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 capacitor 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 capacitor of the example above is charged to 1000 volts, it will discharge to 370 volts in ½ second through the 250,000-ohm resistor.

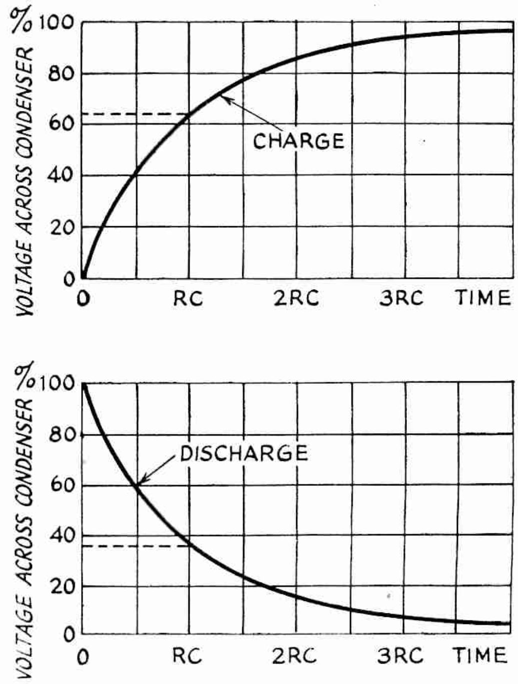


Fig. 2-20—How the voltage across a capacitor rises, with time, when charged through a resistor. The lower curve shows the way in which the voltage decreases across the capacitor terminals on discharging through the same resistor.

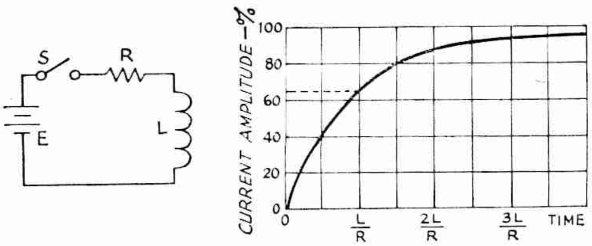


Fig. 2-21—Time constant of an *LR* circuit.

Inductance and Resistance

A comparable situation exists when resistance and inductance are in series. In Fig. 2-21, first consider L to have no 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.

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 growing just fast enough to keep the e.m.f. of self-induction equal to the applied e.m.f.

When resistance is in series, Ohm's Law sets a limit to the value that the current can reach. 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 and so the current never quite reaches the Ohm's Law value, but practically the difference becomes unmeasurable after a time. 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

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

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.

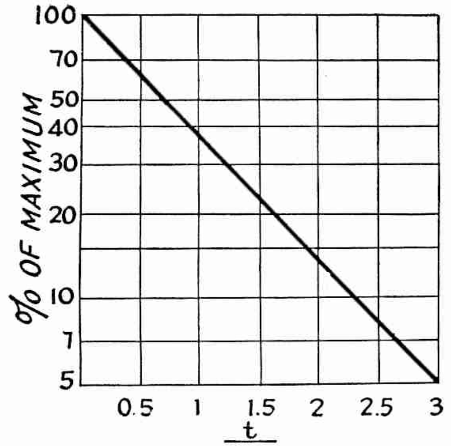


Fig. 2-22—Voltage across capacitor terminals in a discharging *CR* circuit, in terms of the initial charged voltage. To obtain time in seconds, multiply the factor *t/CR* by the time constant of the circuit.

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 capacitor, 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.

Time constants play an important part in numerous devices, such as electronic keys, timing and control circuits, and shaping of keying characteristics by vacuum tubes. The time constants of circuits are also important in such applications as automatic gain control and noise limiters. In nearly all such applications a capacitance-resistance (CR) time constant is involved, and it is usually necessary to know the voltage across the capacitor at some time interval larger or smaller than the actual time constant of the circuit as given by the formula above. Fig. 2-22 can be used for the solution of such problems, since the curve gives the voltage across the capacitor, in terms of percentage of the initial charge, for percentages between 5 and 100, at any time after discharge begins.

Example: A 0.01-μf. capacitor 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.

2—ELECTRICAL LAWS AND CIRCUITS

Alternating Currents

PHASE

The term **phase** essentially means "time," or the *time interval* between the instant when one thing occurs and the instant when a second related thing takes place. The later event is said to **lag** the earlier, while the one that occurs first is said to **lead.** In a.c. circuits the current amplitude changes continuously, so the concept of phase or time becomes important. 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. 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. When two or more frequencies are to be considered, as in the case where harmonics are present, the phase measurements are made with respect to the lowest, or fundamental, frequency.

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 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 is that with 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 instant the cycle began. There is no actual "angle" associated with an alternating current. Fig. 2-23 should help make this method of measurement clear.

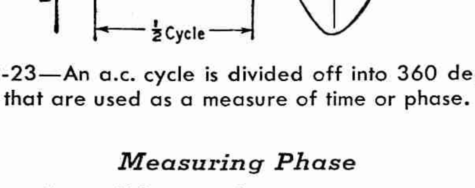


Fig. 2-23—An a.c. cycle is divided off into 360 degrees that are used as a measure of time or phase.

Measuring Phase

The phase difference between two currents of the same frequency is the time or angle difference between corresponding parts of cycles of the two currents. This is shown in Fig. 2-24. The current labeled *A* leads the one marked *B* by 45 degrees, since *A*'s cycles begin 45 degrees earlier in time. It is equally correct to say that *B lags A* by 45 degrees.

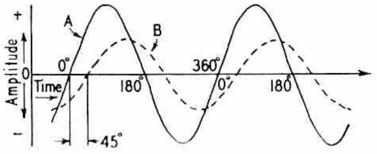


Fig. 2-24—When two waves of the same frequency start their cycles at slightly different times, the time difference or phase difference is measured in degrees. In this drawing wave *B* starts 45 degrees (one-eighth cycle) later than wave *A*, and so lags 45 degrees behind *A*.

Two important special cases are shown in Fig. 2-25. 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.

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

The waves shown in Figs. 2-24 and 2-25 could represent current, voltage, or both. *A* and *B* might be two currents in separate circuits, or *A* might represent voltage and *B* current in the same circuit. If *A* and *B* represent two currents in the *same* circuit (or two voltages in the same circuit) the total or **resultant** current (or voltage) also is a sine wave, because adding any number of sine waves of the same frequency always gives a sine wave also of the same frequency.

Phase in Resistive Circuits

When an alternating voltage is applied to a resistance, the current flows exactly in step with the voltage. In other words, the voltage and current are **in phase.** This is true at any frequency if the resistance is "pure" — that is, is free from the reactive effects discussed in the next section. Practically, it is often difficult to obtain a purely

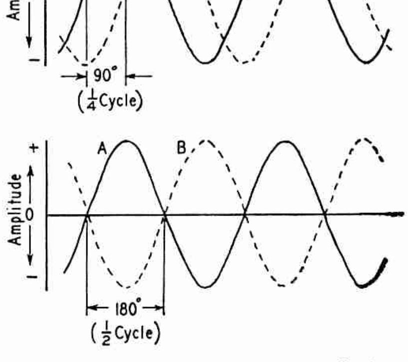


Fig. 2-25—Two important special cases of phase difference. In the upper drawing, the phase difference between A and B is 90 degrees; in the lower drawing the phase difference is 180 degrees.

resistive circuit at radio frequencies, because the reactive effects become more pronounced as the frequency is increased.

In a purely resistive circuit, or for purely resistive parts of circuits, Ohm's Law is just as valid for a.c. of any frequency as it is for d.c.

REACTANCE

Alternating Current in Capacitance

In Fig. 2-26 a sine-wave a.c. voltage having a maximum value of 100 volts is applied to a capacitor. In the period *OA*, the applied voltage increases from zero to 38 volts; at the end of this period the capacitor is charged to that voltage. In interval *AB* the voltage increases to 71 volts; that is, 33 volts additional. In this interval a *smaller* quantity of charge has been added than in *OA*, because the voltage rise during interval *AB* is smaller. Consequently the average current during *AB* is smaller than during *OA*. 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 *AB*, so the quantity of electricity added is less; in other words, the average current during interval *BC* 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 capacitor has the shape of a sine wave, just as the applied voltage does. The current is largest at the beginning of the cycle and becomes zero at the maximum value of the voltage, so there is a phase difference of 90 degrees between the voltage and current. During the first quarter cycle the current is flowing in the normal direction through the circuit, since the capacitor is being charged. Hence the current is positive, as indicated by the dashed line in Fig. 2-26.

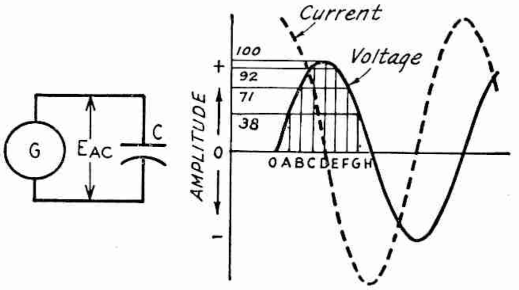


Fig. 2-26—Voltage and current phase relationships when an alternating voltage is applied to a capacitor.

In the second quarter cycle — that is, in the time from *D* to *H*, the voltage applied to the capacitor decreases. During this time the capacitor *loses* its charge. Applying the same reasoning, it is plain that the current is small in interval *DE* and continues to increase during each succeeding interval. However, the current is flowing *against* the applied voltage because the capacitor 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 chaniges to correspond. In other words, an alternateng current flows in the circuit because of the alternate charging and discharging of the capacitance. As shown by Fig. 2-26, the current starts its cycle 90 degrees before the voltage, so the current in a capacitor leads the applied voltage by 90 degrees.

Capacitive Reactance

The quantity of electric charge that can be placed on a capacitor is proportional to the applied e.m.f. and the capacitance. This amount of charge moves back and forth in the circuit once each cycle, and so the *rate* of movement of charge — that is, the current — is proportional to voltage, capacitance and frequency. If the effects of capacitance and frequency are lumped together, they form a quantity that plays a part similar to that of resistance in Ohm's Law. This quantity is called **reactance,** and the unit for it is the ohm, just as in the case of resistance. The formula for it is

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

where X_C = Capacitive reactance in ohms
f = Frequency in cycles per second
C = Capacitance in farads
π = 3.14

Although the unit of reactance is the ohm, there is no power dissipation in reactance. The energy stored in the capacitor in one quarter of the cycle is simply returned to the circuit in the next.

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 capacitor of 470 μμf. (0.00047 μf.) at a frequency of 7150 kc. (7.15 Mc.) is

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

Inductive Reactance

When an alternating voltage is applied to a *pure* inductance (one with no resistance — all *practical* inductors have resistance) the current is again 90 degrees out of phase with the applied voltage. However, in this case the current *lags* 90 degrees behind the voltage — the opposite of the capacitor current-voltage relationship.

The primary cause for this is the *back e.m.f.* generated in the inductance, and since the amplitude of the back e.m.f. is proportional to the rate at which the current changes, and this in turn is proportional to the frequency, the amplitude of the current is inversely proportional to the applied frequency. Also, since the back e.m.f. is proportional to inductance for a given rate of current change, the current flow is inversely proportional to inductance for a given applied voltage and frequency. (Another way of saying this is that just enough current flows to generate an induced e.m.f. that equals and apposes the applied voltage.)

The combined effect of inductance and frequency is called **inductive reactance,** also expressed in ohms, and the formula for it is

$$X_L = 2\pi f L$$

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

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

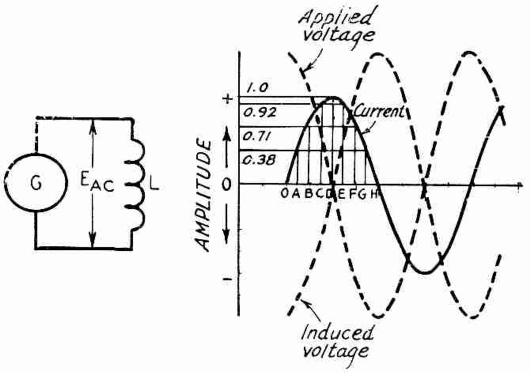


Fig. 2-27 — Phase relationships between voltage and current when an alternating voltage is applied to an inductance.

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

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

The resistance of the wire of which the coil is wound has no effect on the reactance, but simply acts as though it were a separate resistor connected in series with the coil.

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 capacitor of the previous example (reactance = 47.4 ohms) at 7150 kc., the voltage drop across the capacitor is

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

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

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

Reactance Chart

The accompanying chart, Fig. 2-28, shows the reactance of capacitances from 1 $\mu\mu$f. to 100 μf., and the reactance of inductances from 0.1 μh. to 10 henrys, for frequencies between 100 cycles and 100 megacycles per second. The approximate value of reactance can be read from the chart or, where more exact values are needed, the chart will serve as a check on the order of magnitude of reactances calculated from the formulas given above, and thus avoid "decimal-point errors".

Reactances in Series and Parallel

When reactances of the same kind are connected in series or parallel the resultant reactance is that of the resultant inductance or capacitance. This leads to the same rules that are used when determining the resultant resistance when resistors are combined. That is, for series reactances of the same kind the resultant reactance is

$$X = X_1 + X_2 + X_3 + X_4$$

and for reactances of the same kind in parallel the resultant is

$$X = \frac{1}{\frac{1}{X_1} + \frac{1}{X_2} + \frac{1}{X_3} + \frac{1}{X_4}}$$

or for two in parallel,

$$X = \frac{X_1 X_2}{X_1 + X_2}$$

The situation is different when reactances of opposite kinds are combined. Since the current in a capacitance leads the applied voltage by 90 degrees and the current in an inductance lags the applied voltage by 90 degrees, the voltages at the terminals of opposite types of reactance are 180 degrees out of phase in a series circuit (in which the current has to be the same through all elements), and the currents in reactances of opposite types are 180 degrees out of phase in a parallel circuit (in which the same voltage is applied to all elements). The 180-degree phase relationship means that the currents or voltages are of opposite polarity, so in the series circuit of Fig. 2-29A the voltage E_L across the inductive reactance X_L is of opposite polarity to the voltage E_C across the capacitive reactance X_C. Thus if we call X_L "positive" and X_C "negative" (a common convention) the applied voltage E_{AC} is $E_L - E_C$. In

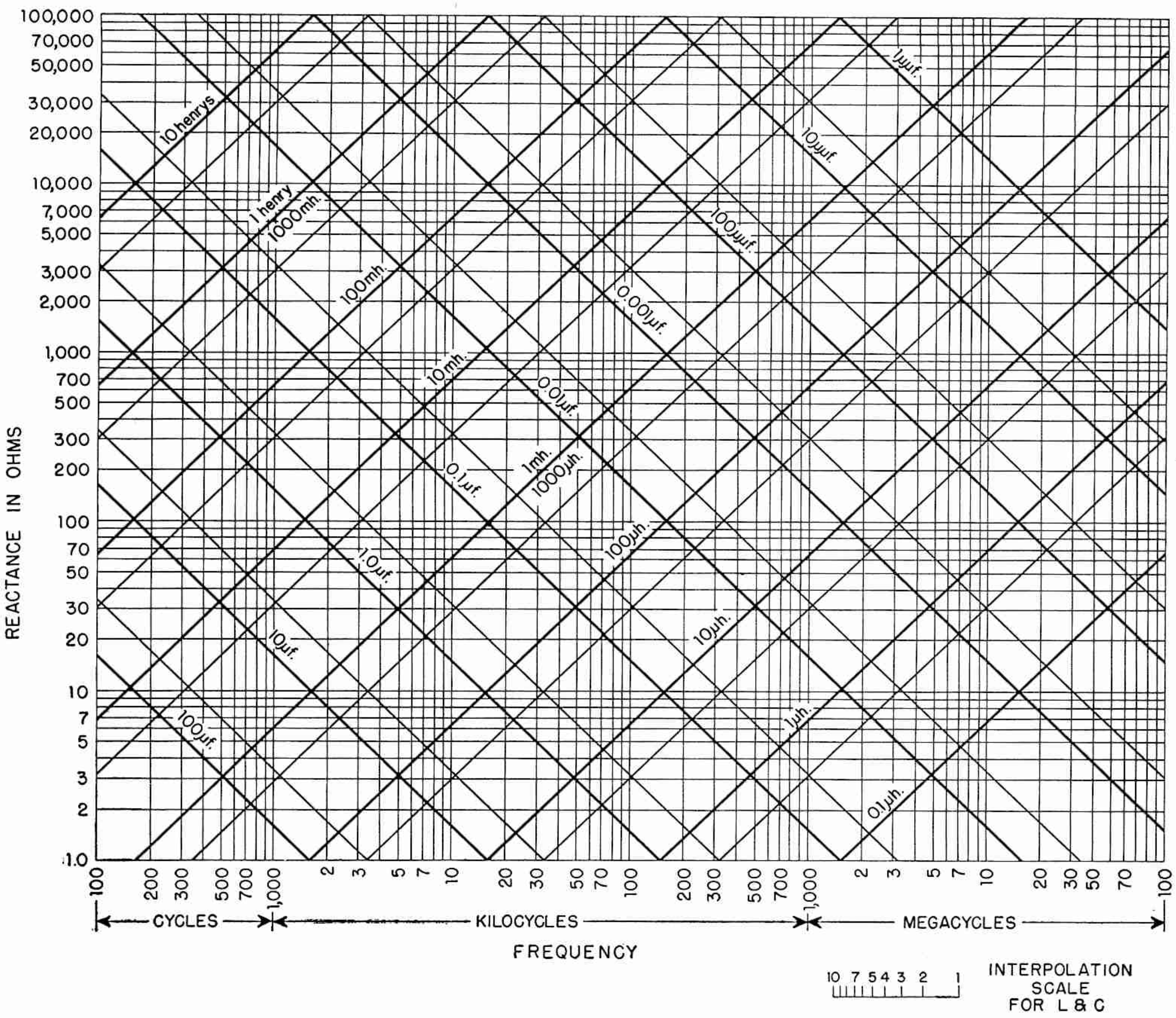


Fig. 2-28—Inductive and capacitive reactance *vs.* frequency. Heavy lines represent multiples of 10, intermediate light lines multiples of 5; e.g., the light line between 10 μh. and 100 μh. represents 50 μh., the light line between 0.1 μf. and 1 μf. represents 0.5 μf., etc. Intermediate values can be estimated with the help of the interpolation scale shown.

Reactances outside the range of the chart may be found by applying appropiate factors to values within the chart range. For example, the reactance of 10 henrys at 60 cycles can be found by taking the reactance of 10 henrys at 600 cycles and dividing by 10 for the 10-times decrease in frequency.

the parallel circuit at B the total current, I, is equal to $I_L - I_C$, since the currents are 180 degrees out of phase.

In the series case, therefore, the resultant reactance of X_L and X_C is

$$X = X_L - X_C$$

and in the parallel case

$$X = \frac{-X_L X_C}{X_L - X_C}$$

Note that in the series circuit the total reactance is negative if X_C is larger than X; this indicates that the total reactance is capacitive in such a case. The resultant reactance in a series circuit is always smaller than the smaller of the two individual reactances.

In the parallel circuit, the resultant reactance is negative (i.e., capacitive) if X_L is larger than X_C, and positive (inductive) if X_L is smaller than X_C, but in every case is always larger than the larger of the two individual reactances.

In the special case where $X_L = X_C$ the total reactance is zero in the series circuit and infinitely large in the parallel circuit.

Reactive Power

In Fig. 2-29A the voltage drop across the inductor is larger than the voltage applied to the circuit. This might seem to be an impossible condition, but it is not; the explanation is that while energy is being stored in the inductor's

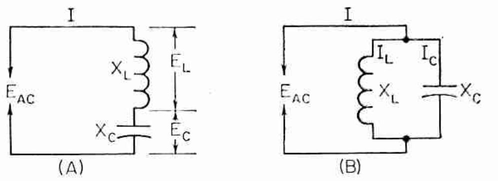


Fig. 2-29—Series and parallel circuits containing opposite kinds of reactance.

magnetic field, energy is being returned to the circuit from the capacitor'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.

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. To distinguish this "nondissipated" power from the power which is actually consumed, the unit of reactive power is called the **volt-ampere-reactive**, or **var**, instead of the watt. Reactive power is sometimes called "wattless" power.

IMPEDANCE

When a circuit contains both resistance and reactance the combined effect of the two is called **impedance**, symbolized by the letter Z. (Impedance is thus a more general term than either resistance or reactance, and is frequently used even for circuits that have only resistance or reactance, although usually with a qualification — such as "resistive impedance" to indicate that the circuit has only resistance, for example.)

The reactance and resistance comprising an impedance may be connected either in series or in parallel, as shown in Fig. 2-30. In these circuits the reactance is shown as a box to indicate that it may be either inductive or capacitive. In the series circuit the current is the same in both elements, with (generally) different voltages appearing across the resistance and reactance. In the parallel circuit the same voltage is applied to both elements, but different currents flow in the two branches.

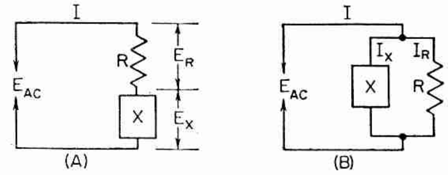


Fig. 2-30—Series and parallel circuits containing resistance and reactance.

Since in a resistance the current is in phase with the applied voltage while in a reactance it is 90 degrees out of phase with the voltage, the phase relationship between current and voltage in the circuit as a whole may be anything between zero and 90 degrees, depending on the relative amounts of resistance and reactance.

Series Circuits

When resistance and reactance are in series, the impedance of the circuit is

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

where Z = impedance in ohms
R = resistance in ohms
X = reactance in ohms.

The reactance may be either capacitive or inductive. If there are two or more reactances in the circuit they may be combined into a resultant by the rules previously given, before substitution into the formula above; similarly for resistances.

The "square root of the sum of the squares" rule for finding impedance in a series circuit arises from the fact that the voltage drops across the resistance and reactance are 90 degrees out of phase, and so combine by the same rule that applies in finding the hypothenuse of a right-angled triangle when the base and altitude are known.

Parallel Circuits

With resistance and reactance in parallel, as in Fig. 2-30B, the impedance is

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

where the symbols have the same meaning as for series circuits.

Just as in the case of series circuits, a number of reactances in parallel should be combined to find the resultant reactance before substitution into the formula above; similarly for a number of resistances in parallel.

Equivalent Series and Parallel Circuits

The two circuits shown in Fig. 2-30 are equivalent if the same current flows when a given voltage of the same frequency is applied, and if the phase angle between voltage and current is the same in both cases. It is in fact possible to "transform" any given series circuit into an equivalent parallel circuit, and vice versa.

Transformations of this type often lead to simplification in the solution of complicated circuits. However, from the standpoint of practical work the usefulness of such transformations lies in the fact that the impedance of a circuit may be modified by the addition of *either* series or parallel elements, depending on which happens to be most convenient in the particular case. Typical applications are considered later in connection with tuned circuits and transmission lines.

Ohm's Law for Impedance

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

Fig. 2-31 shows a simple circuit consisting of a resistance of 75 ohms and a reactance of 100 ohms in series. From the formula previously given, the impedance is

$Z = \sqrt{R^2 + X_L^2} = \sqrt{(75)^2 + (100)^2} = 125$ ohms.

If the applied voltage is 250 volts, then

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

This current flows through both the resistance and reactance, so the voltage drops are

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

The simple arithmetical sum of these two drops, 350 volts, is greater than the applied voltage because the two voltages are 90 degrees out of phase. Their actual resultant, when phase is taken into account, is $\sqrt{(150)^2 + (200)^2} = 250$ volts.

Power Factor

In the circuit of Fig. 2-31 an applied e.m.f. of 250 volts results in a current of 2 amperes, giving an apparent power of $250 \times 2 = 500$ watts. However, only the resistance actually consumes power. The power in the resistance is

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

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

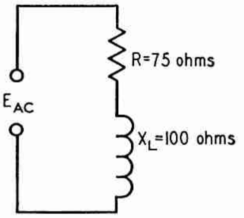


Fig. 2-31—Circuit used as an example for impedance calculations.

"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

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

Reactance and Complex Waves

It was pointed out earlier 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 wave shape as the applied voltage. This is because the reactance of an inductor and capacitor depend upon the applied frequency. For the second-harmonic component of a complex wave, the reactance of the inductor is twice and the reactance of the capacitor one-half their respective values at the fundamental frequency; for the third harmonic the inductor reactance is three times and the capacitor reactance one-third, and so on. Thus the circuit impedance is different for each harmonic component.

Just what happens to the current wave shape 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 harmonic currents will be reduced because the inductive reactance increases in proportion to frequency. When capacitance and resistance are in series, the harmonic current is likely to be accentuated because the capacitive 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 the relative 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

Two coils having mutual inductance constitute a **transformer.** The coil connected to the source of energy is called the **primary** coil, and the other is called the **secondary** coil.

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 a.c. and only a 440-volt source is available, a transformer can be used to change the source voltage to that required. A transformer can be used only with 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.

● THE IRON-CORE TRANSFORMER

As shown in Fig. 2-32, 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

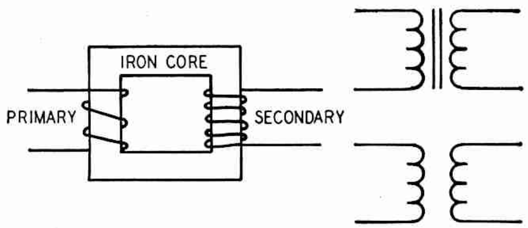


Fig. 2-32—The transformer. Power is transferred from the primary coil to the secondary by means of the magnetic field. The upper symbol at right indicates an iron-core transformer, the lower one an air-core transformer.

having a continuous magnetic path) such as that shown in Fig. 2-32 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 in 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 in each coil. In the primary the induced voltage is practically equal to, and opposes, the applied voltage, as described earlier. Hence,

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

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

The ratio n_s/n_p is called the secondary-to-primary **turns ratio** of the transformer.

Example: A transformer has a primary of 400 turns and a secondary of 2800 turns, and an e.m.f. of 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 an e.m.f. of 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 (enough inductance) 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 with which the primary is wound.

When power is taken from the secondary winding, the secondary current sets up a magnetic field that opposes the field set up by the primary current. But if the induced voltage in the primary is to equal the applied voltage, the original field must be maintained. Consequently, the primary must draw enough additional current to set up a field exactly equal and opposite to the field set up by the secondary current.

In practical calculations on transformers it may be assumed that the entire primary current is caused by the secondary "load." This is justifiable because the magnetizing current should be very small in comparison with the primary "load" current at rated power output.

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

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

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 it and change the e.m.f. 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

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

Transformers

A transformer is usually designed to have its highest efficiency at the 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 by its own losses, because these heat the wire and core. There is a limit to the temperature rise that can be tolerated, because too-high temperature either will melt the wire or cause the insulation to break down. 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** causes an e.m.f. of self-induction; consequently, there are small amounts of **leakage inductance** associated with both windings of the transformer. 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.**

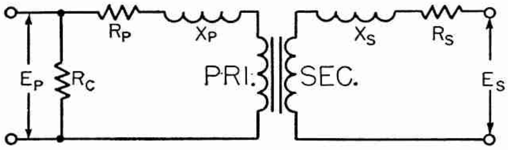


Fig. 2-33—The equivalent circuit of a transformer includes the effects of leakage inductance and resistance of both primary and secondary windings. The resistance R_C is an equivalent resistance representing the core losses, which are essentially constant for any given applied voltage and frequency. Since these are comparatively small, their effect may be neglected in many approximate calculations.

Current flowing through the leakage reactance causes a voltage drop. This voltage drop increases with increasing current, hence it increases as more power is taken from the secondary. Thus, the greater the secondary current, the smaller the secondary terminal voltage becomes. The resistances of the transformer windings 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 looking into primary terminals 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 transformed to a different value "looking into" the primary from the source of power. The 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; thus 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 may be used in practical work 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 requirement is that the primary have enough inductance to operate with low magnetizing current at the voltage applied to the primary.

The primary impedance of a transformer — *as it appears to the source of power* — is determined wholly 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 and frequency at which it is being used. 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 impedance of the actual load that is to dissipate the power may differ widely from this value, so a transformer is used to change the actual load into an impedance of the desired value. This is called **impedance matching.** From the preceding,

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

where N = Required turns ratio, primary to secondary
Z_p = Primary impedance required
Z_s = Impedance of load connected to secondary

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, primary to secondary, required in the coupling transformer is

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

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 deliver 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. 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 and heating from power loss in the source is not important.

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. A short path also helps to reduce flux leakage and therefore minimizes leakage reactance.

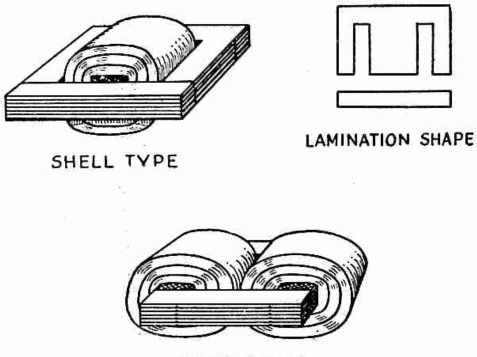


Fig. 2-34—Two common types of transformer construction. Core pieces are interleaved to provide a continuous magnetic path.

Two core shapes are in common use, as shown in Fig. 2-34. 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 capacitive effects between the primary and secondary, or when one of the windings must operate at very high voltage.

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

The number of turns required in the primary for a given applied e.m.f. is determined by the size, shape and type of core material used, and the frequency. The number of turns required is inversely proportional to the cross-sectional area of the core. 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 treated-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. 2-35; 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 will be the case only when the primary (line) and secondary (load) voltages are not very different. The autotransformer is used chiefly for boosting or reducing the power-line voltage by relatively small amounts.

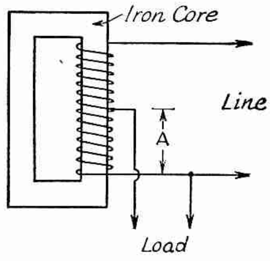


Fig. 2-35—The autotransformer is based on the transformer principle, but uses only one winding. The line and load currents in the common winding (*A*) flow in opposite directions, so that the resultant current is the difference between them. The voltage across A is proportional to the turns ratio.

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 human ear has a *logarithmic* response.

This fact is the basis for the use of the relative-power unit called the **decibel** (abbreviated **db.**) A change of one decibel in the power level is just detectable as a change in loudness under ideal conditions. The number of decibels corresponding to a given power ratio is given by the following formula:

$$Db. = 10 \log \frac{P_2}{P_1}$$

Common logarithms (base 10) are used.

Voltage and Current Ratios

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:

$$Db. = 20 \log \frac{V_2}{V_1}$$

$$\text{or } 20 \log \frac{I_2}{I_1}$$

Decibel Chart

The two formulas are shown graphically in Fig. 2-36 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.

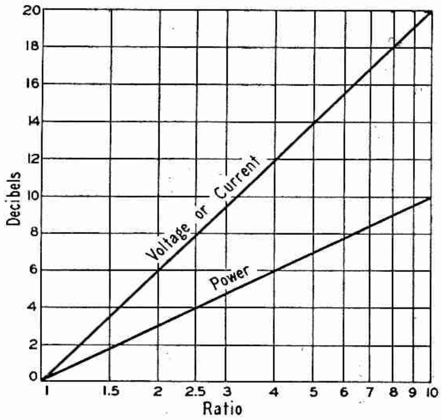


Fig. 2-36—Decibel chart for power, voltage and current ratios for power ratios of 1:1 to 10:1. In determining decibels for current or voltage ratios the currents (or voltages) being compared must be referred to the same value of impedance.

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. For example, a power ratio of 2.5 is 4 db. (from the chart). A power ratio of 10 times 2.5, or 25, is 14 db. (10 + 4), and a power ratio of 100 times 2.5, or 250, is 24 db. (20 + 4). A voltage or current ratio of 4 is 12 db., a voltage or current ratio of 40 is 32 db. (20 + 12), and a voltage or current ratio of 400 is 52 db. (40 + 12).

Radio-Frequency Circuits

● RESONANCE IN SERIES CIRCUITS

Fig. 2-37 shows a resistor, capacitor and inductor connected in series with a source of alternating current, the frequency of which can be varied over a wide range. At some *low* frequency the capacitive 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*. (*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 either of these cases the current will be small, because the reactance is large at either low or high frequencies.

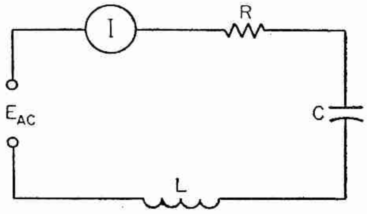


Fig. 2-37—A series circuit containing *L*, *C* and *R* is "resonant" at the applied frequency when the reactance of C is equal to the reactance of *L*.

At some intermediate frequency, the reactances of C and L will be equal and the voltage drops across the coil and capacitor will be equal and 180 degrees out of phase. Therefore they cancel each other completely and the current flow is determined wholly by the resistance, R. At that frequency the current has its largest possible value, assuming 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.**

Although resonance is possible at any frequency, it finds its most extensive application in radio-frequency circuits. The reactive effects associated with even small inductances and capacitances would place drastic limitations on r.f. circuit operation if it were not possible to "cancel them out" by supplying the right amount of reactance of the opposite kind — in other words, "tuning the circuit to resonance."

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 ($\mu\mu$f.)
π = 3.14

Example: The resonant frequency of a series circuit containing a 5-μh. inductor and a 35-$\mu\mu$f. capacitor is

$$= \frac{10^6}{2\pi\sqrt{LC}} = \frac{10^6}{6.28 \times \sqrt{5 \times 35}}$$

$$= \frac{10^6}{6.28 \times 13.2} = \frac{10^6}{83} = 12{,}050 \text{ kc.}$$

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. 2-37 as the frequency is varied (the applied voltage being constant) it would look like one of the curves in Fig. 2-38. The shape of the **resonance curve** at frequencies near resonance is determined by the ratio of reactance to resistance.

If the reactance of either the coil or capacitor 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 frequency moves away from resonance and the circuit is said to be **sharp.** 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.

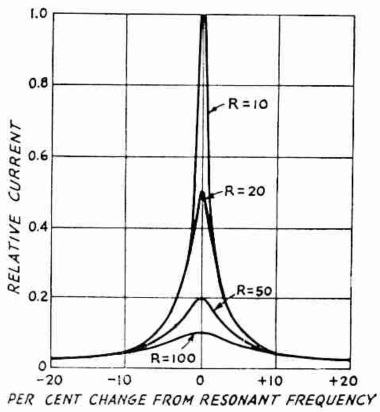


Fig. 2-38—Current in a series-resonant circuit with various values of series resistance. The values are arbitrary and would not apply to all circuits, but represent a typical case. It is assumed that the reactances (at the resonant frequency) are 1000 ohms. Note that at frequencies more than plus or minus ten per cent away from the resonant frequency the current is substantially unaffected by the resistance in the circuit.

Both types of resonance curves are useful. A sharp circuit gives good **selectivity** — the ability to respond strongly (in terms of current amplitude) at 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.

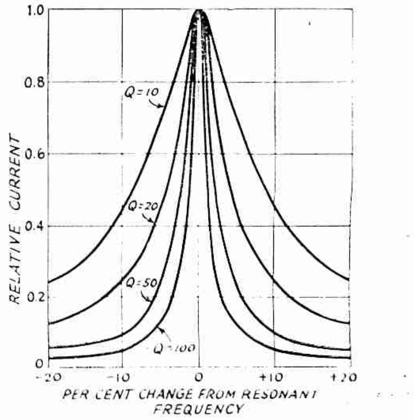


Fig. 2-39—Current in series-resonant circuits having different Qs. In this graph the current at resonance is assumed to be the same in all cases. The lower the Q, the more slowly the current decreases as the applied frequency is moved away from resonance.

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 capacitor (principally in the solid dielectric which must be used to form an insulating support for the capacitor 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.

The value of the reactance of either the inductor or capacitor at the resonant frequency of a series-resonant circuit, 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 = Series resistance in ohms

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

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

The effect of Q on the sharpness of resonance of a circuit is shown by the curves of Fig. 2-39. 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.

Voltage Rise

When a voltage of the resonant frequency is inserted in series in a resonant circuit, the voltage that appears across either the inductor or capacitor is considerably higher than the applied voltage. The current in the circuit is limited only by the resistance and may have a relatively high value; however, the same current flows through the high reactances of the inductor and capacitor and causes large voltage drops. The ratio of the reactive voltage to the applied voltage is equal to the ratio of reactance to resistance. This ratio is also the Q of the circuit. Therefore, the voltage across either the inductor or capacitor 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 inductor or the capacitor 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.

RESONANCE IN PARALLEL CIRCUITS

When a variable-frequency source of constant voltage is applied to a parallel circuit of the type shown in Fig. 2-40 there is a resonance effect similar to that in a series circuit. However, in this case the "line" current (measured at the point indicated) is *smallest* at the frequency for which the inductive and capacitive reactances are equal. At that frequency the current through L is exactly canceled by the out-of-phase current through C, 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.

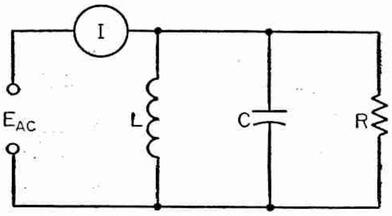
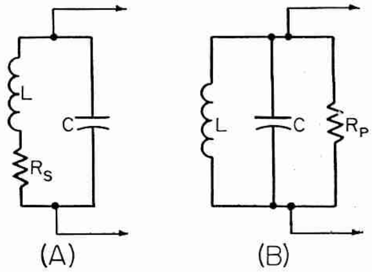


Fig. 2-40—Circuit illustrating parallel resonance.

The resistance R shown in Fig. 2-40 is not necessarily an actual resistor. In most cases it will be an "equivalent" resistance that represents the energy loss in the circuit. This loss can be inherent in the coil or capacitor, 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.)

Parallel and series resonant circuits are quite alike in some respects. For instance, the circuits given at A and B in Fig. 2-41 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



(A) (B)

Fig. 2-41—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.

same Qs. (These statements are approximate, but are quite accurate if the Q is 10 or more.) 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.

Thus a circuit like that of Fig. 2-41A 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 inductive and capacitive currents are 180 degrees out of phase and are equal; thus there is no reactive current in the line. In a practical circuit with a high-Q capacitor, at the resonant frequency the parallel impedance is

$$Z_r = QX$$

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

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

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

At frequencies off resonance the impedance is no longer purely resistive because the inductive and capacitive 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. 2-42 is a set of such curves.

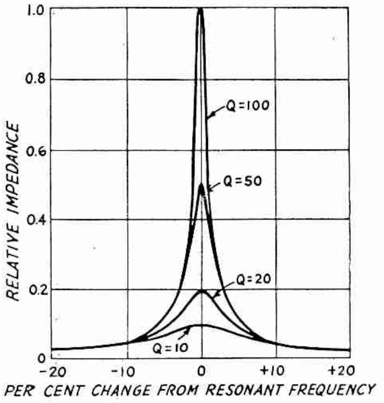


Fig. 2-42—Relative impedance of parallel-resonant circuits with different Qs. These curves are similar to those in Fig. 2-42 for current in a series-resonant circuit. The effect of Q on impedance is most marked near the resonant frequency.

Parallel Resonance in Low-Q Circuits

The preceding discussion is accurate only for Qs of 10 or more. When the Q is below 10, resonance in a parallel circuit having resistance in series with the coil, as in Fig. 2-41A, is not so easily defined. There is a set of values for L and C that will make the parallel impedance a pure resistance, but with these values the impedance does not have its maximum possible value. Another set of values for L and C will make the parallel impedance a maximum, but this maximum value is not a pure resistance. Either condition could be called "resonance," so with low-Q circuits it is necessary to distinguish between **maximum impedance** and **resistive impedance** parallel resonance. The difference between these L and C values and the equal reactances of a series-resonant circuit is appreciable when the Q is in the vicinity of 5, and becomes more marked with still lower Q values.

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 in the coil increases the reactance faster than it raises the resistance, so coils for circuits in which the Q must be high may have reactances of 1000 ohms or more at the frequency under consideration.

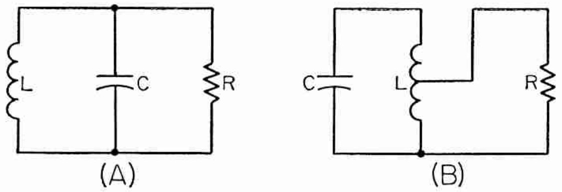


Fig. 2-43—The equivalent circuit of a resonant circuit delivering power to a load. The resistor *R* represents the load resistance. At B the load is tapped across part of *L*, which by transformer action is equivalent to using a higher load resistance across the whole circuit.

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. 2-43A, 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 inductor and capacitor, 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

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

where Q = Quality factor
R = Parallel load resistance (ohms)
X = Reactance (ohms) of either the inductor or capacitor

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

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

The "effective" Q of a circuit loaded by a parallel resistance becomes higher when the reactances 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.

Impedance Transformation

An important application of the parallel-resonant circuit is as an impedance-matching device in the output circuit of a vacuum-tube r.f. power amplifier. As described in the section on vacuum tubes, there is an optimum value of load resistance for each type of tube and set of operating conditions. However, the resistance of the load to which the tube is to deliver power usually is considerably lower than the value required for proper tube operation. To transform the actual load resistance to the desired value the load may be tapped across part of the coil, as shown in Fig. 2-43B. This is equivalent to connecting a higher value of load resistance across the whole circuit, and 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.

When the load resistance has a very low value (say below 100 ohms) it may be connected in series in the resonant circuit (as in Fig. 2-31A, for example), in which case it is transformed to an equivalent parallel impedance as previously described. If the Q is at least 10, the equivalent parallel impedance is

$$Z_r = \frac{X^2}{R}$$

where Z_r = Resistive parallel impedance at resonance

X = Reactance (in ohms) of either the coil or capacitor

R = Load resistance inserted in series

If the Q is lower than 10 the reactance will have to be adjusted somewhat, for the reasons given in the discussion of low-Q circuits, to obtain a resistive impedance of the desired value.

Reactance Values

The charts of Figs. 2-44 and 2-45 show reactance values of inductances and capacitances in the range commonly used in r.f. tuned circuits for the amateur bands. With the exception of the 3.5–4 Mc. band, limiting values for which are shown on the charts, the change in reactance over a band, for either inductors or capacitors, is small enough so that a single curve gives the reactance with sufficient accuracy for most practical purposes.

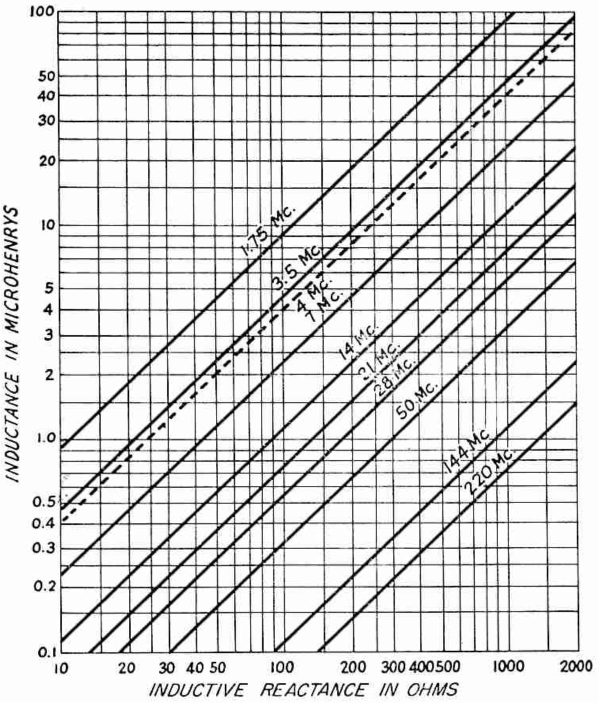


Fig. 2-44—Reactance chart for inductance values commonly used in amateur bands from 1.75 to 220 Mc.

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

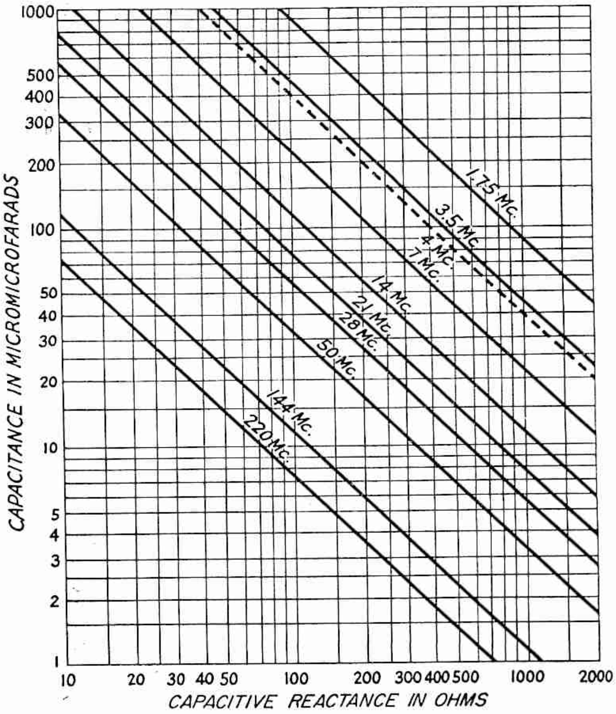


Fig. 2-45—Reactance chart for capacitance values commonly used in amateur bands from 1.75 to 220 Mc.

is one that has more capacitance than "normal" for the frequency; a **low-*C*** circuit one that has less than normal capacitance. These terms depend to a considerable extent upon the particular application considered, and have no exact numerical meaning.

LC Constants

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 ($\mu\mu$f.)
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 $\mu\mu$f. The LC constant is

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

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

COUPLED CIRCUITS

Energy Transfer and Loading

Two circuits are **coupled** when energy can be transferred from one to the other. The circuit delivering power is called the **primary** circuit; the one receiving power is called the **secondary** circuit. The power may be practically all dissipated in the secondary circuit itself (this is usually the case in receiver circuits) or the secondary may simply act as a medium through which the power is transferred to a load. 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.

Coupling by a Common Circuit Element

One method of coupling between two resonant circuits is through a circuit element common to both. The three common variations of this type of coupling are shown in Fig. 2-46; the circuit element common to both circuits carries the subscript M. At A and B current circulating in L_1C_1 flows through the common element, and the voltage developed across this element causes current to flow in L_2C_2. At C, C_M and C_2 form a capacitive voltage divider across L_1C_1, and some of the voltage developed across L_1C_1 is applied across L_2C_2.

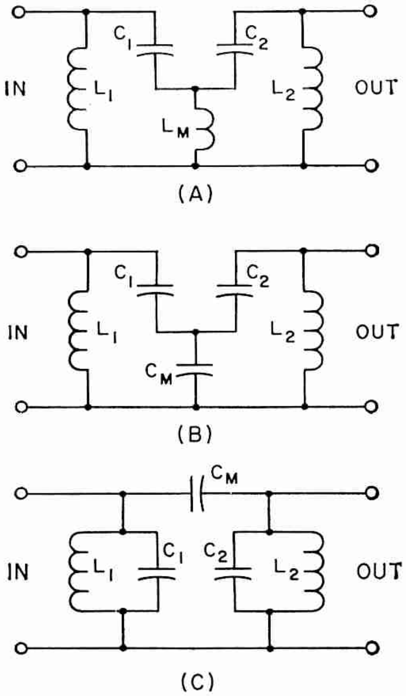


Fig. 2-46—Three methods of circuit coupling.

If both circuits are resonant to the same frequency, as is usually the case, the value of coupling reactance required for maximum energy transfer can be approximated by the following, based on $L_1 = L_2$, $C_1 = C_2$ and $Q_1 = Q_2$:

(A) $L_M \approx L_1/Q_1$; (B) $C_M \approx Q_1C_1$; (C) $C_M \approx C_1/Q_1$.

The coupling can be increased by increasing the above coupling elements in A and C and decreasing the value in B. When the coupling is increased, the resultant bandwidth of the combination is increased, and this principle is sometimes applied to "broad-band" the circuits in a transmitter or receiver. When the coupling elements in A and C are decreased, or when the coupling element in B is increased, the coupling between the circuits is decreased below the *critical coupling* value on which the above approximations are based. Less than critical coupling will decrease the bandwidth and the energy transfer; the principle is often used in receivers to improve the selectivity.

Inductive Coupling

Figs. 2-47 and 2-48 show inductive coupling, or coupling by means of the mutual inductance between two coils. Circuits of this type resemble the iron-core transformer, but because only a part of

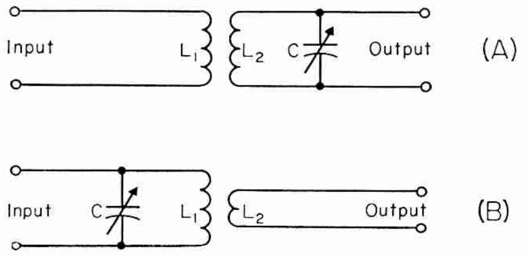


Fig. 2-47—Single-tuned inductively coupled circuits.

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.

Two types of inductively-coupled circuits are shown in Fig. 2-47. Only one circuit 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.

In these circuits the coupling between the primary and secondary coils usually is "tight" — that is, the coefficient of coupling between the coils is large. With very tight coupling either circuit operates nearly 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, thus either circuit is approximately equivalent to Fig. 2-43B.

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. In any case, the maximum energy transfer possible for a given coefficient of coupling is obtained when the reactance of the untuned coil is equal to the resistance of its load.

The *Q* and parallel impedance of the tuned circuit are reduced by coupling through an untuned coil in much the same way as by the tapping arrangement shown in Fig. 2-43B.

Coupled Resonant Circuits

When the primary and secondary circuits are both tuned, as in Fig. 2-48, 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 primary circuit is connected to a source of r.f. energy of the resonant frequency and the secondary is then loosely coupled to the primary, a current will flow in the secondary circuit. In flowing through the resistance of the secondary circuit and any load that may be connected to it, the current causes a power loss. This power must come from the energy source through the primary circuit, and manifests itself in the primary as an increase in the equivalent resistance in series with the primary coil. Hence the *Q* and parallel impedance of the primary circuit are decreased by the coupled secondary. 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 to a maximum at one value of coupling, called **critical coupling,** but then decreases if the coupling is tightened still more (still without changing the tuning).

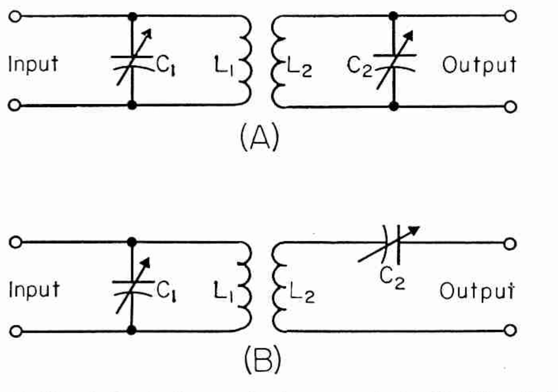


Fig. 2-48—Inductively-coupled resonant circuits. Circuit A is used for high-resistance loads (load resistance much higher than the reactance of either L_2 or C_2 at the resonant frequency). Circuit B is suitable for low resistance loads (load resistance much lower than the reactance of either L_2 or C_2 at the resonant frequency).

Critical coupling is a function of the *Q*s of the two circuits. 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 such as are used in transmitters the *Q* may be too low to give the desired power transfer even when the coils are coupled as tightly as the physical construction permits. In such case, increasing the *Q* of either circuit will be helpful, although it is generally better to increase the *Q* of the lower-*Q* circuit rather than the reverse. The *Q* of the parallel-tuned primary (input) circuit can be increased by decreasing the *L/C* ratio because, as shown in connection with Fig. 2-43, this circuit is in effect loaded by a parallel resistance (effect of coupled-in resistance). In the parallel-tuned secondary circuit, Fig. 2-48A, the *Q* can be increased, for a fixed value of load resistance, either by decreasing the *L/C* ratio or by tapping the load down (see Fig. 2-43). In the series-tuned secondary circuit, Fig. 2-48B, the *Q* may be increased by *increasing* the *L/C* ratio. There will generally be no difficulty in securing sufficient coupling, with practicable coils, if the product of the *Q*s of the two tuned circuits is 10 or more. A smaller product will suffice if the coil construction permits tight coupling.

Selectivity

In Fig. 2-47 only one circuit is tuned and the selectivity curve will be essentially that of a single resonant circuit. As stated, the effective *Q* depends upon the resistance connected to the untuned coil.

In Fig. 2-48, 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 (this is not the condition for optimum power transfer discussed immediately above) and both circuits are tuned to resonance. The *Q*s of the individual circuits

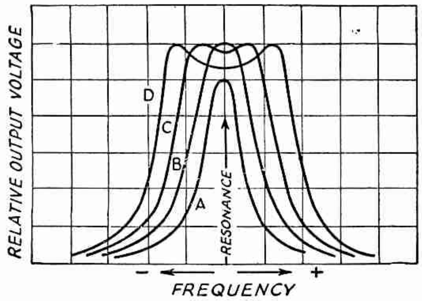


Fig. 2-49—Showing the effect on the output voltage from the secondary circuit of changing the coefficient of coupling between two resonant circuits independently tuned to the same frequency. The voltage applied to the primary is held constant in amplitude while the frequency is varied, and the output voltage is measured across the secondary.

are affected by the degree of coupling, because each couples resistance into the other; the tighter the coupling, the lower the individual Qs and therefore the lower the over-all selectivity.

If both circuits are independently tuned to resonance, the over-all selectivity will vary about as shown in Fig. 2-49 as the coupling is varied. With 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. 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 each hump represents a new condition of critical coupling at a frequency to which the primary is tuned by the additional coupled-in reactance from the secondary.

Band-Pass Coupling

Over-coupled resonant circuits are useful where substantially uniform output is desired over a continuous band of frequencies, without readjustment of tuning. The width of the flat top of the resonance curve depends on the Qs of the two circuits as well as the tightness of coupling; the frequency separation between the humps will increase, and the curve become more flat-topped, as the Qs are lowered.

Band-pass operation also is secured by tuning the two circuits to slightly different frequencies, which gives a double-humped resonance curve even with loose coupling. This is called **stagger tuning.** However, to secure adequate power transfer over the frequency band it is usually necessary to use tight coupling and experimentally adjust the circuits for the desired performance.

Link Coupling

A modification of inductive coupling, called **link coupling,** is shown in Fig. 2-50. 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.

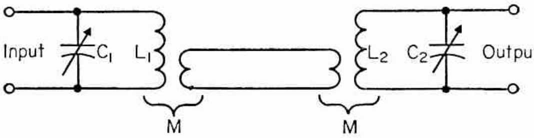


Fig. 2-50—Link coupling. The mutual inductances at both ends of the link are equivalent to mutual inductance between the tuned circuits, and serve the same purpose.

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* inductance. The length of the link between the coils is not critical if it is very small compared with the wavelength, but if the length is more than about one-twentieth 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.

● IMPEDANCE-MATCHING CIRCUITS

The coupling circuits discussed in the preceding section have been based either on inductive coupling or on coupling through a common circuit element between two resonant circuits. These are not the only circuits that may be used for transferring power from one device to another. There is, in fact, a wide variety of such circuits available, all of them being classified generally as **impedance-matching networks.** Several networks frequently used in amateur equipment are shown in Fig. 2-51.

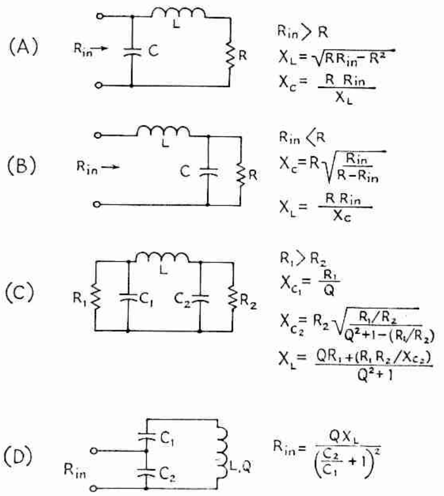


Fig. 2-51—Impedance-matching networks adaptable to amateur work. (A) L network for transforming to a higher value of resistance. (B) L network for transforming to a lower resistance value. (C) Pi network. R_1 is the larger of the two resistors; Q is defined as R_1/X_{C1}. (D) Tapped tuned circuit used in some receiver applications. The impedance of the tuned circuit is transformed to a lower value, R_{in}, by the capacitive divider.

The L Network

The L network is the simplest possible impedance-matching circuit. It closely resembles an ordinary resonant circuit with the load resistance, R, Fig. 2-51, either in series or parallel. The arrangement shown in Fig. 2-51A is used when the desired impedance, R_{IN}, is larger than the actual load resistance, R, while Fig. 2-51B is used in the opposite case. The design equations for each case are given in the figure, in terms of the circuit reactances. The reactances may be converted to inductance and capacitance by means of the formulas previously given or taken directly from the charts of Figs. 2-44 and 2-45.

When the impedance transformation ratio is large — that is, one of the two impedances is of the order of 100 times (or more) larger than the other — the operation of the circuit is exactly the same as previously discussed in connection with impedance transformation with a simple *LC* resonant circuit.

The Q of an L network is found in the same way as for simple resonant circuits. That is, it is equal to X_L/R or R_{IN}/X_C in Fig. 2-51A, and to X_L/R_{IN} or R/X_C in Fig. 2-51B. The value of Q is determined by the ratio of the impedances to be matched, and cannot be selected independently. In the equations of Fig. 2-51 it is assumed that both R and R_{IN} are pure resistances.

The Pi Network

The pi network, shown in Fig. 2-51C, offers more flexibility than the L since the operating Q may be chosen practically at will. The only limitation on the circuit values that may be used is that the reactance of the series arm, the inductor L in the figure, must not be greater than the square root of the product of the two values of resistive impedance to be matched. As the circuit is applied in amateur equipment, this limiting value of reactance would represent a network with an undesirably low operating Q, and the circuit values ordinarily used are well on the safe side of the limiting values.

In its principal application as a "tank" circuit matching a transmission line to a power amplifier tube, the load R_2 will generally have a fairly low value of resistance (up to a few hundred ohms) while R_1, the required load for the tube, will be of the order of a few thousand ohms. In such a case the Q of the circuit is defined as R_1/X_{C1}, so the choice of a value for the operating Q immediately sets the value of X_{C1} and hence of C_1. The values of X_{C2} and X_L are then found from the equations given in the figure.

Graphical solutions of these equations for the most important practical cases are given in the chapter on transmitter design in the discussion of plate tank circuits. The L and C values may be calculated from the reactances or read from the charts of Figs. 2-44 and 2-45.

Tapped Tuned Circuit

The tapped tuned circuit of Fig. 2-51D is useful in some receiver applications, where it is desirable to use a high-impedance tuned circuit as a lower-impedance load. When the Q of the inductor has been determined, the capacitors can be selected to give the desired impedance transformation and the necessary resultant capacitance to tune the circuit to resonance.

● FILTERS

A **filter** is an electrical circuit configuration **(network)** designed to have specific characteristics with respect to the transmission or attenuation of various frequencies that may be applied to it. There are three general types of filters: **low-pass, high-pass,** and **band-pass.**

A low-pass filter is one that will permit all frequencies below a specified one called the **cut-off frequency** to be transmitted with little or no loss, but that will attenuate all frequencies above the cut-off frequency.

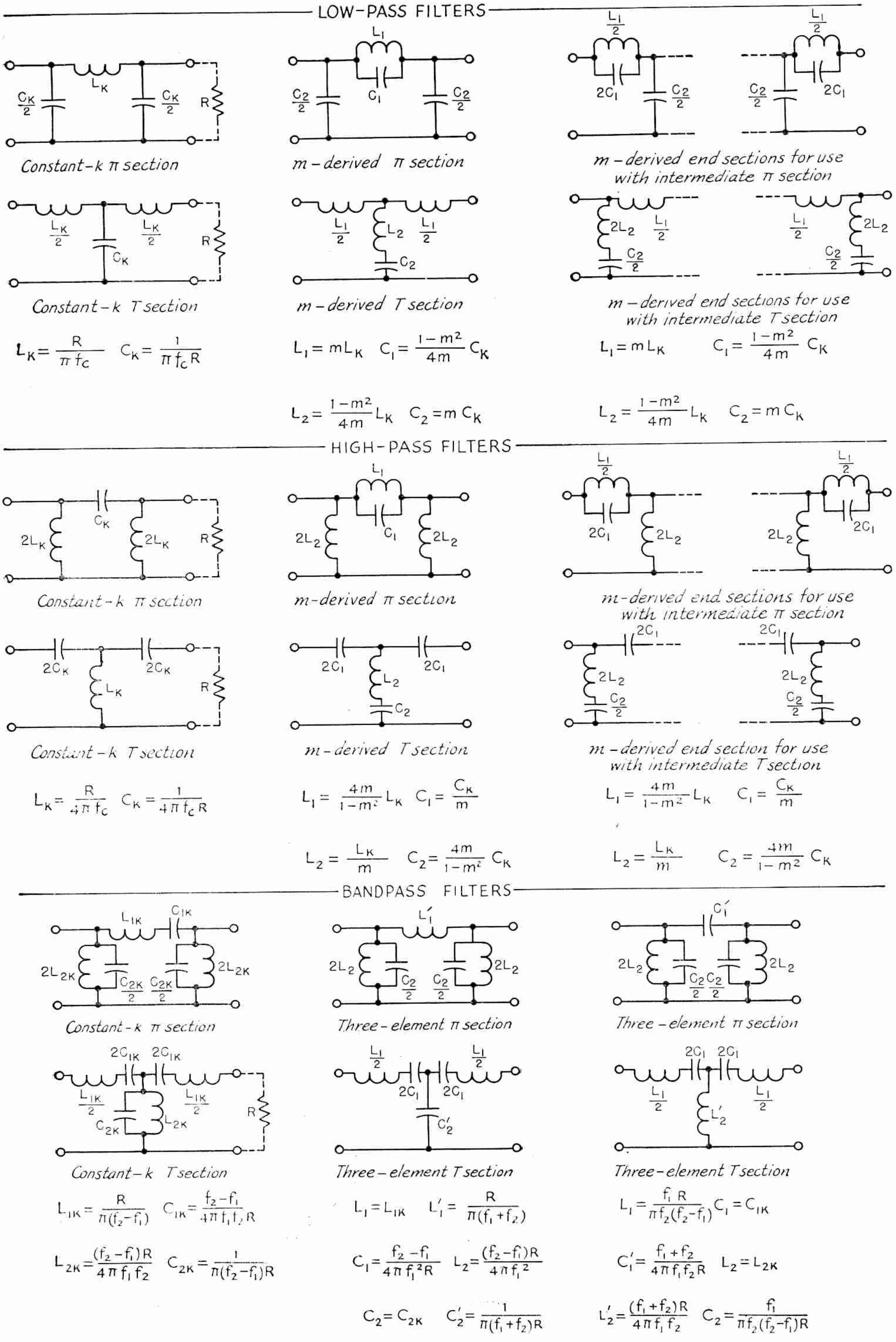
A high-pass filter similarly has a cut-off frequency, above which there is little or no loss in transmission, but below which there is considerable attenuation. Its behavior is the opposite of that of the low-pass filter.

A band-pass filter is one that will transmit a selected band of frequencies with substantially no loss, but that will attenuate all frequencies either higher or lower than the desired band.

The **pass band** of a filter is the frequency spectrum that is transmitted with little or no loss. The transmission characteristic is not necessarily perfectly uniform in the pass band, but the variations usually are small.

The **stop band** is the frequency region in which attenuation is desired. The attenuation may vary in the stop band, and in a simple filter usually is least near the cut-off frequency, rising to high values at frequencies considerably removed from the cut-off frequency.

Filters are designed for a specific value of purely resistive impedance (the **terminating im-**



LOW-PASS FILTERS

Constant-k π section — *Constant-k T section*

$L_K = \frac{R}{\pi f_c}$ $C_K = \frac{1}{\pi f_c R}$

m-derived π section — *m-derived T section*

$L_1 = mL_K$ $C_1 = \frac{1-m^2}{4m} C_K$

$L_2 = \frac{1-m^2}{4m} L_K$ $C_2 = m\,C_K$

m-derived end sections for use with intermediate π section — *m-derived end sections for use with intermediate T section*

$L_1 = mL_K$ $C_1 = \frac{1-m^2}{4m} C_K$

$L_2 = \frac{1-m^2}{4m} L_K$ $C_2 = m\,C_K$

HIGH-PASS FILTERS

Constant-k π section — *Constant-k T section*

$L_K = \frac{R}{4\pi f_c}$ $C_K = \frac{1}{4\pi f_c R}$

m-derived π section — *m-derived T section*

$L_1 = \frac{4m}{1-m^2} L_K$ $C_1 = \frac{C_K}{m}$

$L_2 = \frac{L_K}{m}$ $C_2 = \frac{4m}{1-m^2} C_K$

m-derived end sections for use with intermediate π section — *m-derived end section for use with intermediate T section*

$L_1 = \frac{4m}{1-m^2} L_K$ $C_1 = \frac{C_K}{m}$

$L_2 = \frac{L_K}{m}$ $C_2 = \frac{4m}{1-m^2} C_K$

BANDPASS FILTERS

Constant-k π section — *Constant-k T section*

$L_{1K} = \frac{R}{\pi(f_2-f_1)}$ $C_{1K} = \frac{f_2-f_1}{4\pi f_1 f_2 R}$

$L_{2K} = \frac{(f_2-f_1)R}{4\pi f_1 f_2}$ $C_{2K} = \frac{1}{\pi(f_2-f_1)R}$

Three-element π section — *Three-element T section*

$L_1 = L_{1K}$ $L_1' = \frac{R}{\pi(f_1+f_2)}$

$C_1 = \frac{f_2-f_1}{4\pi f_1^2 R}$ $L_2 = \frac{(f_2-f_1)R}{4\pi f_1^2}$

$C_2 = C_{2K}$ $C_2' = \frac{1}{\pi(f_1+f_2)R}$

Three-element π section — *Three-element T section*

$L_1 = \frac{f_1 R}{\pi f_2(f_2-f_1)}$ $C_1 = C_{1K}$

$C_1' = \frac{f_1+f_2}{4\pi f_1 f_2 R}$ $L_2 = L_{2K}$

$L_2' = \frac{(f_1+f_2)R}{4\pi f_1 f_2}$ $C_2 = \frac{f_1}{\pi f_2(f_2-f_1)R}$

Fig. 2-52—Basic filter sections and design formulas. In the above formulas *R* is in ohms, C in farads, *L* in henrys, and *f* in cycles per second.

pedance of the filter). When such an impedance is connected to the output terminals of the filter, the impedance looking into the input terminals has essentially the same value, throughout most of the pass band. Simple filters do not give perfectly uniform performance in this respect, but the input impedance of a properly-terminated filter can be made fairly constant, as well as closer to the design value, over the pass band by using **m-derived** filter sections.

A discussion of filter design principles is beyond the scope of this *Handbook*, but it is not difficult to build satisfactory filters from the circuits and formulas given in Fig. 2-52. Filter circuits are built up from elementary sections as shown in the figure. These sections can be used alone or, if greater attenuation and sharper cut-off (that is, a more rapid rate of rise of attenuation with frequency beyond the cut-off frequency) 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 band-pass 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 per second, respectively.

All of the types shown are "unbalanced" (one side grounded). For use in balanced circuits (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 inductors of a value equal to $L_k/2$, while the balanced constant-k π-section high-pass filter would use two capacitors each 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 center section, although an m-derived center section can be used. The factor m determines the ratio of the cut-off frequency, f_c, to a frequency of high attenuation, f_∞. 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 output sides of the filters shown should be terminated in a resistance equal to R, and there should be little or no reactive component in the termination.

Simple audio filters can be made with powdered-iron-core inductors and paper capacitors. 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 inductors on toroidial powdered-iron forms; three sections are generally needed for obtaining equivalent results when using other types of inductors.

Band-pass filters for single sideband work (see later chapter) are often 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 is the most suitable capacitor dielectric.

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 capacitors, depending upon the power requirements.

In any filter, there should be no magnetic or capacitive 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.

● 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 the **piezoelectric effect.** A small plate or bar cut in the proper way from a quartz crystal 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 be developed between the electrodes.

Piezoelectric crystals can be used to transform mechanical energy into electrical energy, and vice versa. They are used 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. Crystals of Rochelle salts are used for these purposes.

Crystal Resonators

Crystalline plates also are mechanical resonators that have natural frequencies of vibration ranging from a few thousand cycles to tens of 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. The thing that makes the **crystal resonator** valuable is that it has extremely high Q, ranging from 5 to 10 times the Qs *obtainable* with good LC resonant circuits.

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 holder plates between which it is sandwiched; these plates form, with the crystal as the dielectric, a small capacitor like any other capacitor constructed of two plates with a dielectric between. The crystal itself is equivalent to a series-resonant circuit, and together with the capacitance of the holder forms the equivalent circuit shown in Fig. 2-53. At frequencies of the order of 450 kc., where crystals are widely used as resonators, the equivalent L may be several henrys and the equivalent C only a few hundredths of a micromicrofarad. Although the equivalent R is of the order of a few thousand ohms, the reactance at resonance is so high that the Q of the crystal likewise is high.

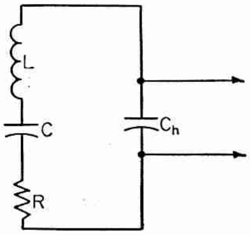


Fig. 2-53—Equivalent circuit of a crystal resonator. *L, C* and *R* are the electrical equivalents of mechanical properties of the crystal; C_h is the capacitance of the holder plates with the crystal plate between them.

A circuit of the type shown in Fig. 2-53 has a series-resonant frequency, when viewed from the circuit terminals indicated by the arrowheads, determined by L and C only. At this frequency the circuit impedance is simply equal to R, providing the reactance of C_h is large compared with R (this is generally the case). The circuit also has a parallel-resonant frequency determined by L and the equivalent capacitance of C and C_h in series. Since this equivalent capacitance is smaller than C alone, the parallel-resonant frequency is higher than the series-resonant frequency. The separation between the two resonant frequencies depends on the ratio of C_h to C, and when this ratio is large (as in the case of a crystal resonator, where C_h will be a few μμf. in the average case) the two frequencies will be quite close together. A separation of a kilocycle or less at 455 kc. is typical of a quartz crystal.

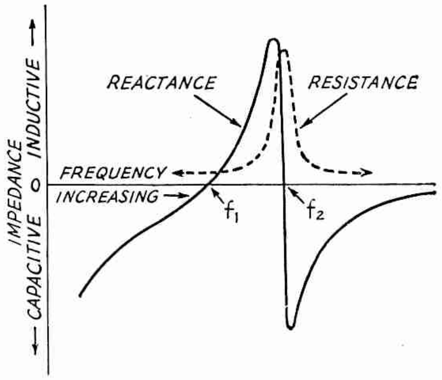


Fig. 2-54—Reactance and resistance *vs.* frequency of a circuit of the type shown in Fig. 2-53. Actual values of reactance, resistance and the separation between the series- and parallel-resonant frequencies, f_1, and f_2, respectively, depend on the circuit constants.

Fig. 2-54 shows how the resistance and reactance of such a circuit vary as the applied frequency is varied. The reactance passes through zero at both resonant frequencies, but the resistance rises to a large value at parallel resonance, just as in any tuned circuit.

Quartz crystals may be used either as simple resonators for their selective properties or as the frequency-controlling elements in oscillators as described in later chapters. The series-resonant frequency is the one principally used in the former case, while the more common forms of oscillator circuit use the parallel-resonant frequency.

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 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. 2-55. It is convenient to consider that the alternating current is **superimposed** on the direct current, so we may look upon the actual current as having two components, one d.c. and the other a.c.

In an alternating current the positive and negative alternations have the same average amplitude, so 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 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 varies as the square of the instantaneous value of the current, and when all the instantaneous squared values are averaged over a cycle the total power is greater than the d.c. power alone. If the a.c. is a

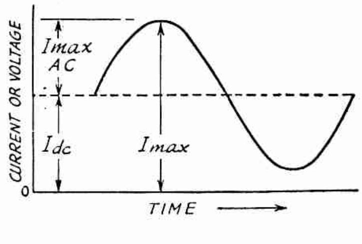
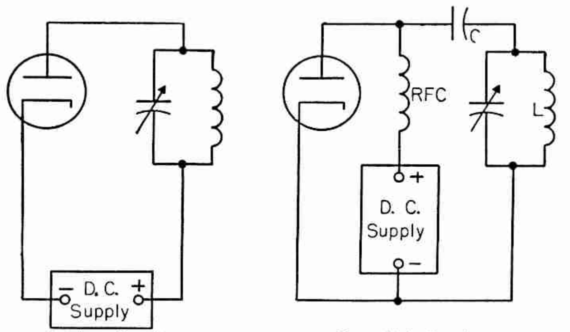


Fig. 2-55—Pulsating d.c., composed of an alternating current or voltage superimposed on a steady direct current or voltage.

Practical Circuit Details



Series Feed

Parallel Feed

Fig. 2-56—Illustrating series and parallel feed.

sine wave having a peak value just equal to the d.c., the power in the circuit is 1.5 times the d.c. power. An instrument whose readings are proportional to power will show such an increase.

Series and Parallel Feed

Fig. 2-56 shows in simplified form how d.c. and a.c. may be combined in a vacuum-tube circuit. In this case, it is assumed that the a.c. is at radio frequency, as suggested by the coil-and-capacitor tuned circuit. It is also assumed 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. coil 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 capacitance,** *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.**

Either type of feed 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, a desirable feature from the viewpoint of safety to the operator, because the voltages applied to tubes — particularly transmitting tubes — are dangerous. 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 often preferred, therefore, because it is relatively easy to keep the impedance between the a.c. circuit and the tube low.

Bypassing

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, partly because 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.

An actual circuit would be provided with a **bypass capacitor,** as shown in Fig. 2-57. Capacitor *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. Hence the r.f. current will tend to flow through it rather than through the d.c. supply.

To be effective, the reactance of the bypass capacitor should not be more than one-tenth of the impedance of the bypassed 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 bypass 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. 2-57.

The same type of bypassing is used when audio frequencies are present in addition to r.f. Because the reactance of a capacitor 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 µf. 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.) Bypass capacitors also are used in audio circuits to carry the audio frequencies around a d.c. supply.

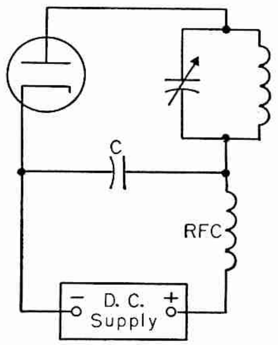


Fig. 2-57—Typical use of a bypass capacitor in a series-feed circuit.

Distributed Capacitance and Inductance

In the discussions earlier in this chapter it was assumed that a capacitor has only capacitance and that an inductor 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 capacitor 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 capacitor and the **distributed capacitance** in an inductor have important practical effects. Actually, every capacitor 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 capacitor will act like a normal capacitance and the coil will act like a normal inductance. Near the natural resonant points, the coil and capacitor act like self-tuned circuits. Above resonance, the capacitor acts like an inductor and the inductor acts like a capacitor. 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 bypassing and choking as well. It will be appreciated that a bypass capacitor that actually acts like an inductance, or an r.f. choke that acts like a low-reactance capacitor, cannot work as it is intended they should.

Grounds

Throughout this book there are frequent references to **ground** and **ground potential.** When a connection is said to be "grounded" it does not necessarily mean that it actually goes to earth. What it means is that an actual earth connection to that point in the circuit should not disturb 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 general practice, 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, and 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. 2-58. R.f. circuits are shown in the upper row, while iron-core transformers (such as are used in power-supply and audio circuits) are shown in the lower row. 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" capacitor and connecting its 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.

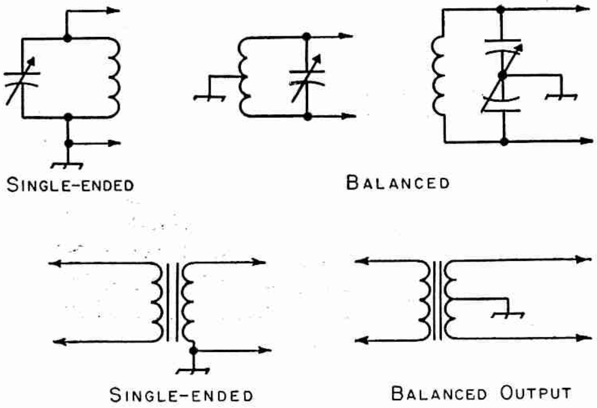


Fig. 2-58—Single-ended and balanced circuits.

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 for magnetic fields increases with frequency and with the conductivity and thickness of the shielding material.

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

Shielding a coil reduces its inductance, because part of its field is canceled by the shield. 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, but the reduction in inductance and Q will be small if the spacing between the sides of the coil and the shield is at least half the coil diameter, and if the spacing at the ends of the coil is at least equal to 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.

U.H.F. Circuits

● RESONANT LINES

In resonant circuits as employed at the lower frequencies it is possible to consider each of the reactance components as a separate entity. The fact that an inductor has a certain amount of self-capacitance, as well as some resistance, while a capacitor also possesses a small self-inductance, can usually be disregarded.

At the very-high and ultrahigh frequencies it is not readily possible to separate these components. Also, the connecting leads, which at lower frequencies would serve merely to join the capacitor and 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.

For these reasons it is common practice to utilize resonant sections of transmission line as tuned circuits at frequencies above 100 Mc. or so. A quarter-wavelength line, or any odd multiple thereof, shorted at one end and open at the other exhibits large standing waves, as described in the section on transmission lines. 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 The equivalent relationships are shown in Fig. 2-59. At frequencies off resonance the line displays qualities comparable with the inductive and capacitive reactances of a conventional tuned circuit, so sections of transmission line can be used in much the same manner as inductors and capacitors.

To minimize radiation loss the two conductors of a parallel-conductor line should not be more than about one-tenth wavelength apart, the spacing being measured between the conductor axes. On the other hand, the spacing should not be less than about twice the conductor diameter because of "proximity effect," which causes eddy currents and an increase in loss. Above 300 Mc. it is difficult to satisfy both these requirements simultaneously, and the radiation from an open line tends to become excessive, reducing the Q. In such case the coaxial type of line is to be preferred, since it is inherently shielded.

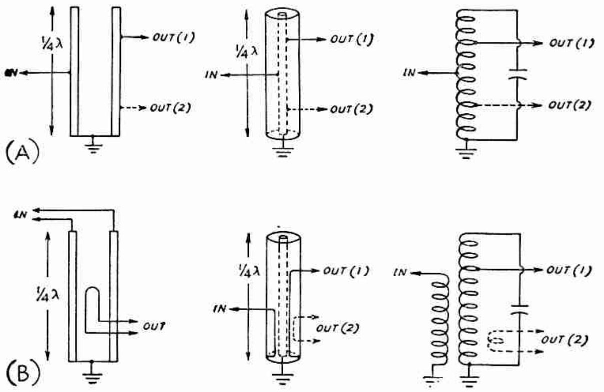


Fig. 2-59—Equivalent coupling circuits for parallel-line, coaxial-line and conventional resonant circuits.

Representative methods for adjusting coaxial lines to resonance are shown in Fig. 2-60. 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 using parallel-plate capacitors are illustrated. The arrangement with the loading capacitor at the open

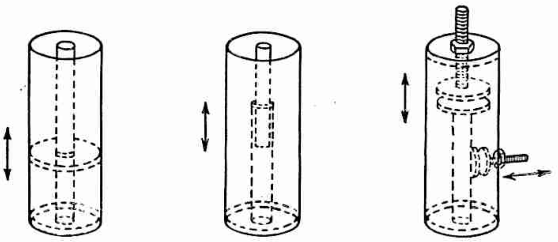


Fig. 2-60—Methods of tuning coaxial resonant lines.

end of the line has the greatest tuning effect per unit of capacitance; the alternative method, which is equivalent to tapping the capacitor 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 unloaded lines resonant at the same frequency.

Two methods of tuning a parallel-conductor lines are shown in Fig. 2-61. The sliding short-circuiting strap can be tightened by means of screws and nuts to make good electrical contact. The parallel-plate capacitor in the second drawing may be placed anywhere along the line, the tuning effect becoming less as the capacitor is located nearer the shorted end of the line. Although a low-capacitance variable capacitor 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.

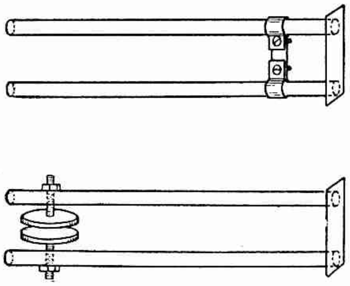


Fig. 2-61 —Methods of tuning parallel-type resonant lines.

WAVEGUIDES

A waveguide 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 waveguide 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.

Analysis of waveguide 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. 2-62. It will be observed that the intensity of the electric field is greatest (as indicated by closer spacing of the lines of force) at the center along the x dimension, Fig. 2-62B, 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.

Modes of Propagation

Fig. 2-62 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.

Waveguide Dimensions

In the rectangular guide the critical dimension is x in Fig. 2-62; 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 $\frac{1}{2}x$

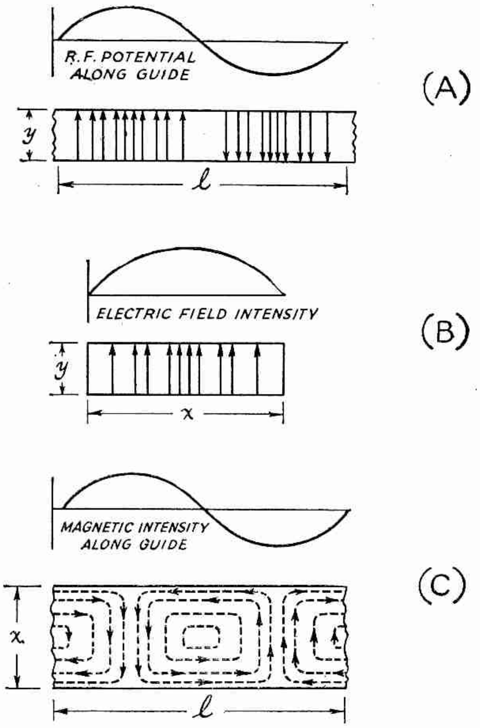


Fig. 2-62—Field distribution in a rectangular waveguide. The $TE_{1,0}$ mode of propagation is depicted

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

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 waveguide with the dimensions chosen so that waves of a given length can be maintained inside.

Typical shapes used for resonators are the cylinder, the rectangular box and the sphere, as shown in Fig. 2-63. 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 waveguide). For the lowest modes the resonant wavelengths are as follows:

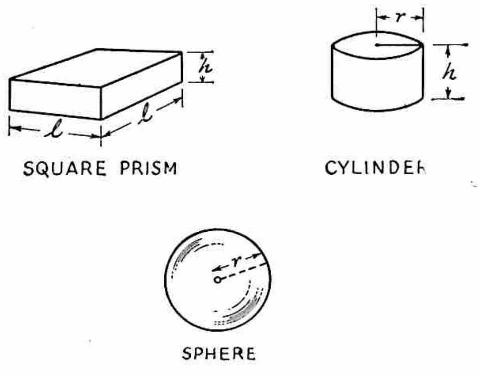


Fig. 2-63—Forms of cavity resonators.

Cylinder	$2.61r$
Square box	$1.41l$
Sphere	$2.28r$

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. A cylindrical cavity can be tuned by a sliding shorting disk 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.

A form of cavity resonator in practical use is the re-entrant cylindrical type shown in Fig. 2-64. In construction it resembles a concentric line closed at both ends with capacitive 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.

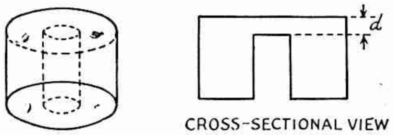


Fig. 2-64—Re-entrant cylindrical cavity resonator.

Compared with 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 be secured with good design and construction.

Coupling to Waveguides and Cavity Resonators

Energy may be introduced into or abstracted from a waveguide 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. 2-65. 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.

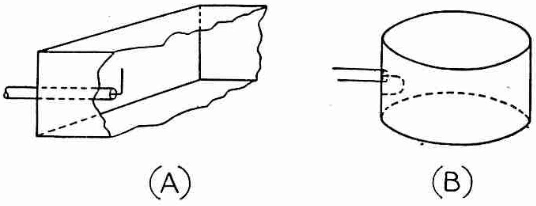


Fig. 2-65—Coupling to waveguides and resonators.

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.

Modulation, Heterodyning and Beats

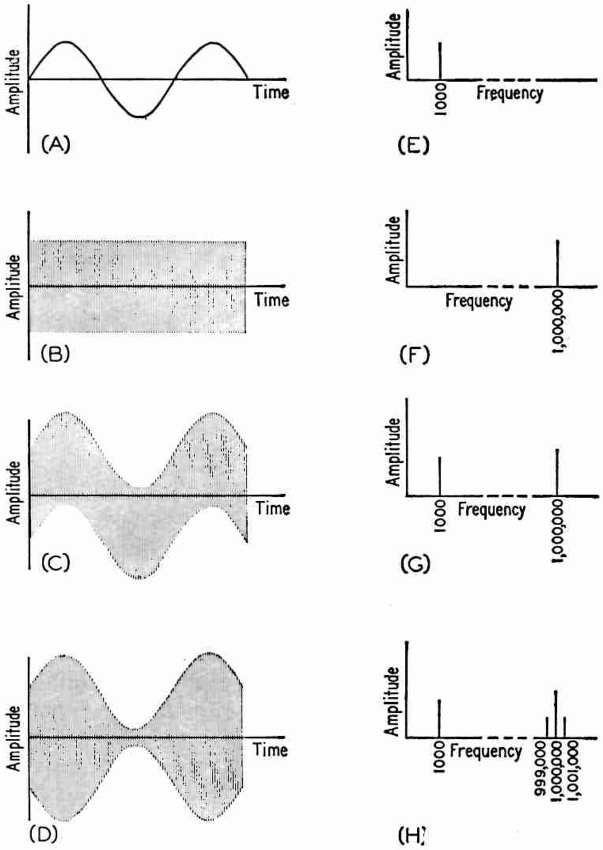


Fig. 2-66—Amplitude-*vs.*-time and amplitude-*vs.*-frequency plots of various signals. (A) 1½ cycles of an audio signal, assumed to be 1000 c.p.s. in this example. (B) A radio-frequency signal, assumed to be 1 Mc. (1,000,000 c.p.s.); 1500 cycles are completed during the same time as the 1½ cycles in A, so they cannot be shown accurately. (C) The signals of A and B in the same circuit; each maintains its own identity. (D) The signals of A and B in a circuit where the amplitude of A can control the amplitude of B. The 1-Mc. signal is *modulated* by the 1000-cycle signal. E, F, G, and H show amplitude-*vs.*-frequency plots of the signals in A, B, C and D, respectively. Note the new frequencies in H, resulting from the modulation process.

Since one of the most widespread uses of radio frequencies is the 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. Suppose the audio signal to be transmitted by radio is a pure 1000-cycle tone, and we wish to transmit it at 1 Mc. (1,000,000 cycles). One possible way might be to add 1,000,000 cycles and 1,000 cycles together, thereby obtaining a radio frequency of 1,001,000 cycles. No simple method for doing such a thing directly has ever been devised, although the *effect* is obtained and used in advanced communications techniques.

Actually, when two different frequencies are present simultaneously in an ordinary circuit (specifically, one in which Ohm's Law holds) each behaves as though the other were not there. It is true that the total or resultant voltage (or current) in the circuit will be the sum of the instantaneous values of the two at every instant. This is because there can be only one value of current or voltage at any single point in a circuit at any instant. Figs. 2-66A and B show two such frequencies, and C shows the resultant. The amplitude of the 1,000,000-cycle current is not affected by the presence of the 1000-cycle current, but merely has its axis shifted back and forth at the 1000-cycle rate. An attempt to transmit such a combination as a radio wave would result simply in the transmission of the 1,000,000-cycle frequency, since the 1000-cycle frequency retains its identity as an audio frequency and hence will not be radiated.

There are devices, however, which make it possible for one frequency to control the amplitude of the other. If, for example, a 1000-cycle tone is used to control a 1-Mc. signal, the maximum r.f. output will be obtained when the 1000-cycle signal is at the peak of one alternation and the minimum will occur at the peak of the next alternation. The process is called **amplitude modulation,** and the effect is shown in Fig. 2-66D. The resultant signal is now entirely at radio frequency, but with its amplitude varying at the modulation rate (1000 cycles). Receiving equipment adjusted to receive the 1,000,000-cycle r.f. signal can reproduce these changes in amplitude, and thus tell what the audio signal is, through a process called **detection** or **demodulation.**

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) of the two, and hence the radio frequencies appearing in the circuit after modulation are 999,000, 1,000,000 and 1,001,000 cycles.

When an audio frequency is used to control the amplitude of a radio frequency, the process is generally called "amplitude modulation," as mentioned, but when a radio frequency modulates another radio frequency it is called **heterodyning.** However, the processes are identical. 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.**

Modulation, Heterodyning and Beats

In A, B, C and D of Fig. 2-66, 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. 2-66 show the signals of Fig. 2-66A, B, C and D on an amplitude-*vs.*-frequency basis. Any one frequency is, of course, represented by a vertical line. Fig. 2-66H shows the side frequencies appearing as a result of the modulation process.

Amplitude modulation **(a.m.)** is not the only possible type nor is it the only one in use. Any signal property can be modulated. These properties include frequency and phase as well as amplitude, and methods are available for modulating all three. However, in every case the modulation process leads to the generation of a new set of radio frequencies symmetrically disposed about the original radio frequency (**carrier** frequency). The various types of modulation are treated in detail in later chapters.

CHAPTER 3

Vacuum-Tube Principles

● CURRENT IN A VACUUM

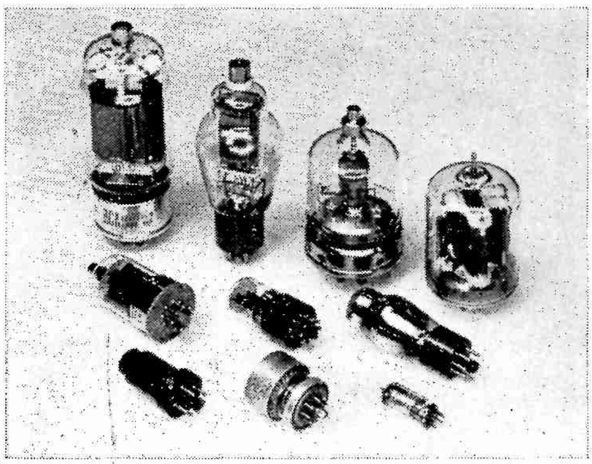
The outstanding difference between the vacuum tube and most other electrical devices is that the electric current does not flow through a conductor but through empty space — a vacuum. This is only possible when "free" electrons — that is, electrons that are not attached to atoms — are somehow introduced into the vacuum. Free electrons in an evacuated space will be attracted to a positively charged object within the same space, or will 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 space charge repels those electrons nearest the cathode, tending to make them fall back on it.



Representative tube types. Transmitting tubes having up to 500-watt capability are shown in the back row. The tube with the top cap in the middle row is a low-power transmitting type. Others are receiving tubes, with the exception of the one in the center foreground which is a v.h.f. transmitting type.

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 by connecting a source of e.m.f. between it and the cathode, as indicated in Fig. 3-1, electrons emitted by the cathode are attracted to the positively charged conductor. An electric current then flows through the circuit formed by the cathode, the charged conductor, and 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.

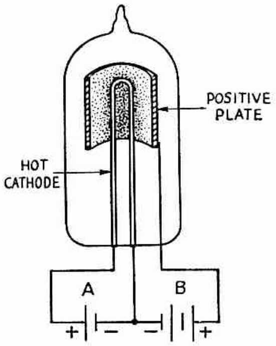


Fig. 3-1 — Conduction by thermionic emission in a vacuum tube. The A battery is used to heat the filament to a temperature that will cause it to emit electrons. The B battery makes the plate positive with respect to the filament, thereby causing the emitted electrons to be attracted to the plate. Electrons captured by the plate flow back through the B 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. 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. However, it is not essential that the heating cur-

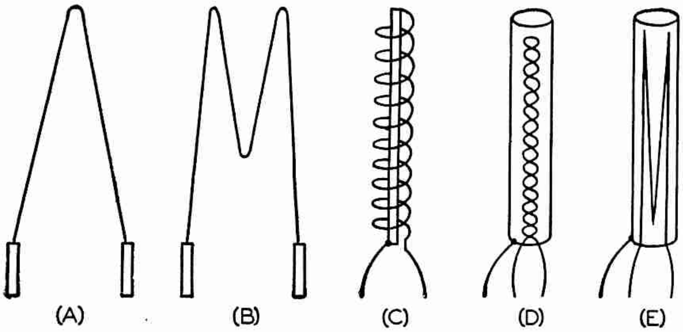


Fig. 3-2—Types of cathode construction. Directly heated cathodes or filaments are shown at A, B, and C. The inverted V filament is used in small receiving tubes, the M in both receiving and transmitting tubes. The spiral filament is a transmitting-tube type. The indirectly-heated cathodes at D and E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

rent flow through the actual material that does the emitting; the filament or heater can be electrically separate from the emitting cathode. 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.

Much greater electron emission can be obtained, at relatively low temperatures, by using special cathode materials rather than pure metals. 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.

Plate Current

If there is only a small positive voltage on the plate, the number of electrons reaching it will be small because the space charge (which is negative) prevents those electrons nearest the cathode from being attracted to the plate. As the plate voltage is increased, the effect of the space charge is increasingly overcome and the number of electrons attracted to the plate becomes larger. That is, the **plate current** increases with increasing plate voltage.

Fig. 3-3 shows a typical plot of plate current *vs.* 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 is 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 until a **saturation point** is reached. This is where the positive charge on the plate has substantially overcome the space charge and almost all the electrons are going to the plate. At higher voltages the plate current stays at practically the same value.

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.

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 load resistor, *R*, represents the actual circuit in which the rectified alternating current does work. All tubes work with 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 cause power to be developed in 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.

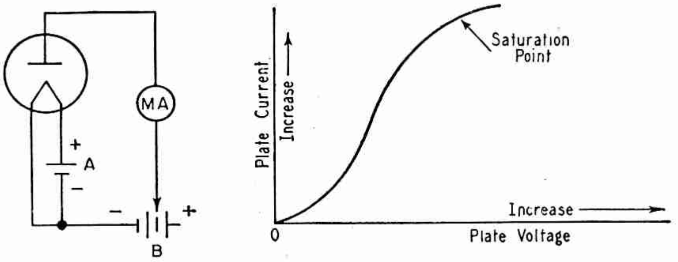
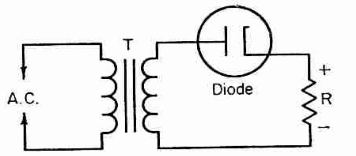


Fig. 3-3—The diode, or two-element tube, and a typical curve showing how the plate current depends upon the voltage applied to the plate.

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.



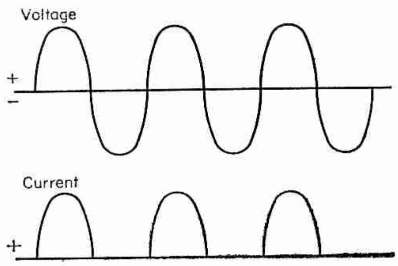


Fig. 3-4—Rectification in a diode. Current flows only when the plate is positive with respect to the cathode, so that only half-cycles of current flow through the load resistor, *R*.

Vacuum-Tube Amplifiers

TRIODES

Grid Control

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.

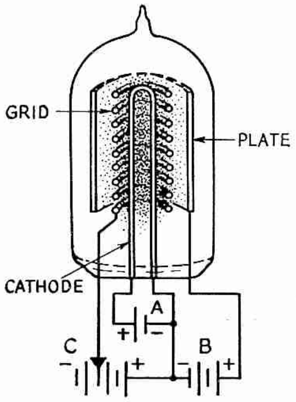


Fig. 3-5—Construction of an elementary triode vacuum tube, showing the filament, grid (with an end view of the grid wires) and plate. The relative density of the space charge is indicated roughly by the dot density.

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 to reach 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. 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.

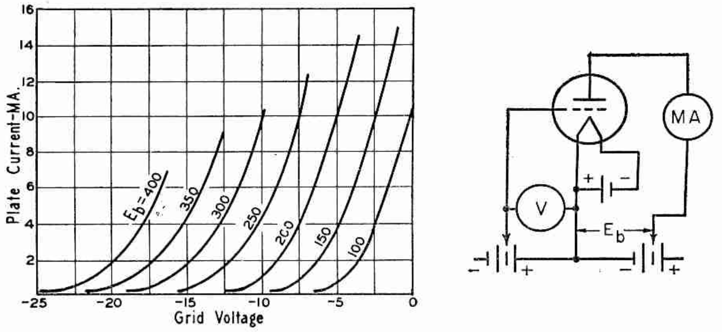


Fig. 3-6—Grid-voltage-vs.-plate-current curves at various fixed values of plate voltage (E_b) for a typical small triode. Characteristic curves of this type can be taken by varying the battery voltages in the circuit at the right.

The curves could be extended by making the grid voltage positive as well as negative. When the grid is negative, it repels electrons and therefore none of them reaches it; in other words, no current flows in the grid circuit. However, 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 no power is used.

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

The fact that a small voltage acting on the grid is equivalent to a large voltage acting on the plate indicates the possibility of **amplification** with the triode tube. The many uses of the electronic tube nearly all are based upon this amplifying feature. The amplified output 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 that 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 100 or so. 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 to obtain a high μ it is necessary to construct the grid with many turns of wire per inch, or in the form of a fine mesh. This leaves a relatively small open area for electrons to go through to reach the plate, so 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.

The best all-around indication of the effectiveness of the tube as an amplifier is its **grid-plate transconductance** — also called **mutual conductance.** This characteristic takes account of both amplification factor and plate resistance, and therefore is a figure of merit for the tube. 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 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

The way in which a tube amplifies is best shown by 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 in Fig. 3-7 a load resistance is connected in series with the plate of the tube. 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.

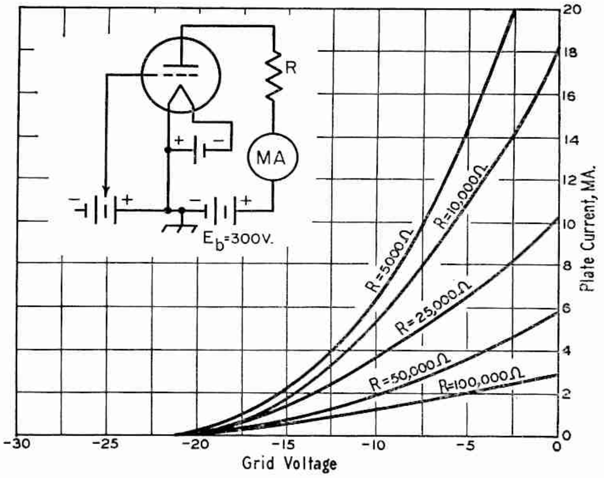


Fig. 3-7—Dynamic characteristics of a small triode with various load resistances from 5000 to 100,000 ohms.

The several curves in Fig. 3-7 are for various values of load resistance. 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; also, the curve tends to be straighter.

Fig. 3-8 is 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.

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

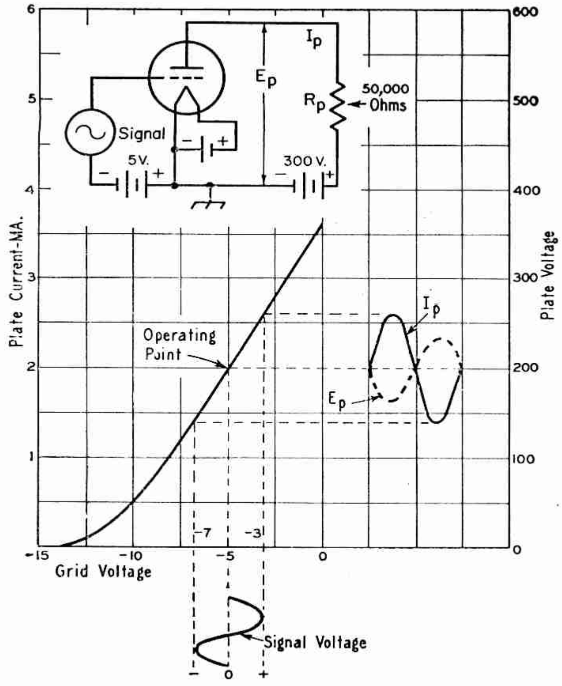


Fig. 3-8—Amplifier operation. When the plate current varies in response to the signal applied to the grid, a varying voltage drop appears across the load, R_p, as shown by the dashed curve, E_p. I_p is the plate current.

and cathode of the tube also is shown on the 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.

As shown by the drawings in Fig. 3-8, 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 voltage 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. One object of the type of amplification shown in this drawing is to obtain, from the plate circuit, an alternating voltage that has the same wave-shape as the signal voltage applied to the grid. To do so, an **operating point** on the straight part of the curve must be selected. 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 wave shape will be **distorted.**

A second reason for using negative grid bias is that 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 peak of the signal does exceed the negative bias, current will flow in the grid circuit during the time the grid is positive.)

Distortion of the output wave shape 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

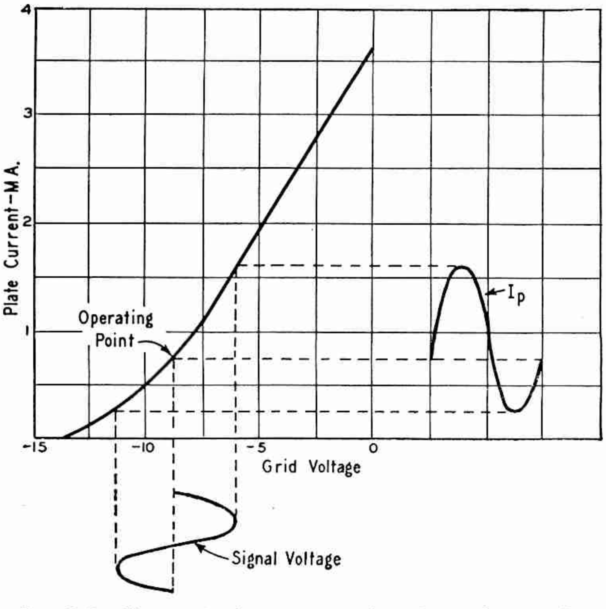


Fig. 3-9—Harmonic distortion resulting from choice of an operating point on the curved part of the tube characteristic. The lower half-cycle of plate current does not have the same shape as the upper half-cycle.

there are occasions when harmonics are deliberately generated and used.

Amplifier Output Circuits

The useful output 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 for the tube's operation, but 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.

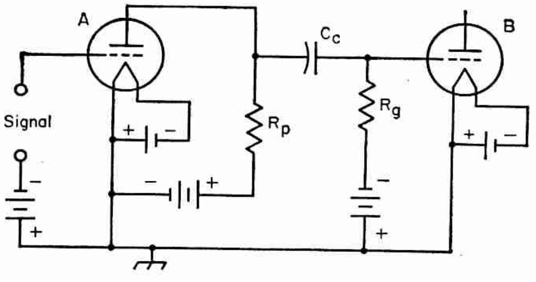
Three types of coupling are in common use at audio frequencies. These are **resistance coupling, impedance coupling,** and **transformer coupling.** They are shown in Fig. 3-10. In all three cases the output is shown coupled to the grid circuit of a subsequent amplifier tube, but the same types of circuits 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, the a.c. voltage between the plate and cathode of the tube) is applied to a second resistor, R_g, through a **coupling capacitor,** C_c. The capacitor "blocks off" the d.c. voltage on the plate of the first tube and prevents it from being applied to the grid of tube *B*. The latter tube has negative grid bias 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*.

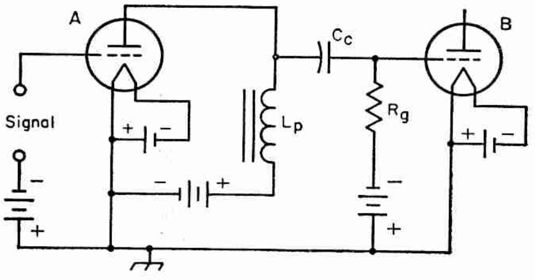
The **grid resistor,** R_g, usually has a rather high value (0.5 to 2 megohms). The reactance of the coupling capacitor, C_c, must be low enough compared with the resistance of R_g so that the a.c. voltage drop in C_c is negligible at the lowest frequency to be amplified. If R_g is at least 0.5 megohm,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

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.

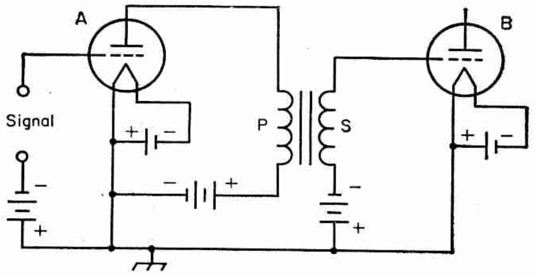
The impedance-coupled circuit differs from that using resistance coupling only in the substitution of a high-inductance coil (usually several hundred henrys for audio frequencies) for the plate resistor. The advantage of using an inductance rather than a resistor is that its impedance is high for alternating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



RESISTANCE COUPLING



IMPEDANCE COUPLING



TRANSFORMER COUPLING

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

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 d.c. voltage from the plate supply. Also, if the secondary has more turns than the primary, the output voltage will be "stepped up" in proportion to the turns ratio.

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 in voltage amplifiers (see below) is best suited to triodes having amplification factors of about 20 or less, for the reason that the primary inductance of a practicable transformer cannot be made large enough to work well with a tube having high plate resistance.

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

Voltage amplifiers belong to a group called **Class A amplifiers.** A Class A amplifier is one operated so that the wave shape 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. For 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.

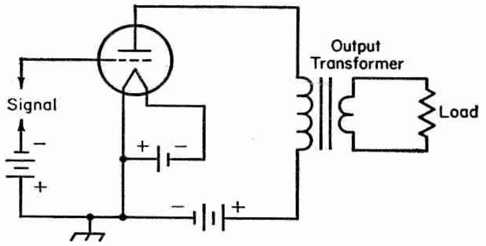


Fig. 3-11—An elementary power-amplifier circuit in which the power-consuming load is coupled to the plate circuit through an impedance-matching transformer.

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. Every power tube requires a specific value of load resistance from plate to cathode, usually some thousands of ohms, for optimum operation. The resistance of the actual load is rarely the right value for "matching" this optimum load resistance, so 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 wave shape.

Another term used in connection with power 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.

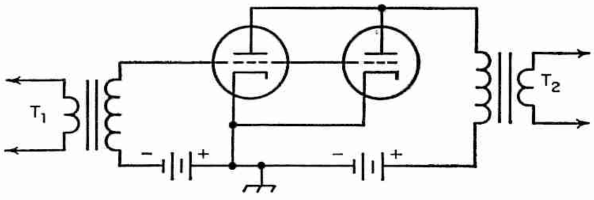
The a.c. power that is delivered to a load by an amplifier tube has to be paid for in power taken from the source of plate voltage and current. In fact, there is always more power going into the plate circuit of the tube than is coming out as useful output. The difference between the input and output power is used up in heating the plate of the tube, as explained previously. 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 given plate-dissipation rating.

Parallel and Push-Pull

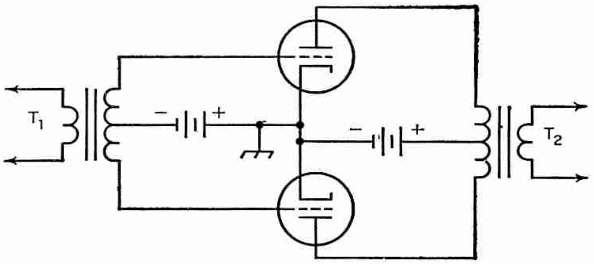
When it is necessary to obtain more power output than one tube is capable of giving, two or more similar tubes may be connected in **parallel.** In this case the similar elements in all tubes are connected together. This method is shown in Fig. 3-12 for a transformer-coupled amplifier. The power output 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 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.



PARALLEL



PUSH-PULL

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

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 readily 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 stages used successively are said to be in **cascade.**

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 with respect to the cathode. Since in the balanced grid circuit the signal voltages on the grids of the two tubes always have opposite polarities, 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 has considerably higher plate efficiency 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 d.c. plate input that can be applied to a Class A amplifier is equal to 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.

A Class B amplifier usually is operated in such a way as to secure the maximum possible power output. This requires rather large values of plate current, and to obtain them the signal voltage must completely overcome the grid bias 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 preceding amplifier or **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 B amplifiers used at radio frequencies are known as **linear amplifiers** because they are

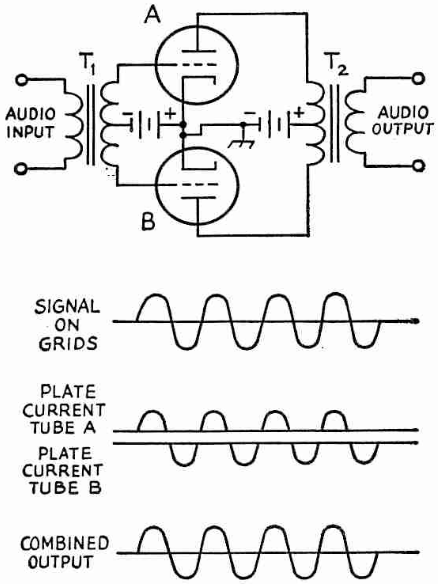


Fig. 3-13—Class B amplifier operation.

adjusted to operate in such a way that the power output is proportional to the square of the r.f. exciting voltage. This permits amplification of a modulated r.f. signal without distortion. Push-pull is not required in this type of operation; a single tube can be used equally well.

Class AB Amplifiers

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 if 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 that will deliver power, without distortion, into a load of highly variable resistance.

Operating Angle

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 leads 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 merely put together two distorted half-cycles. An operating angle of less than 180 degrees therefore cannot be used if distortionless output is wanted.

Class C Amplifiers

In power amplifiers operating at radio frequencies distortion of the r.f. wave form is relatively unimportant. For reasons described later in this chapter, an r.f. amplifier must be operated with tuned circuits, and the selectivity of such circuits "filters out" the r.f. harmonics resulting from distortion.

A radio-frequency power amplifier therefore can be used with an operating angle of less than 180 degrees. This is called **Class C** operation. The advantage is that the plate efficiency is increased, because the loss in the plate is proportional, among other things, to the amount of time during which the plate current flows, and this time is reduced by decreasing the operating angle.

Depending on the type of tube, the optimum load resistance for a Class C amplifier ranges from about 1500 to 5000 ohms. It is usually secured by using tuned-circuit arrangements, of the type described in the chapter on circuit fundamentals, to transform the resistance of the actual load to the value required by the tube. The grid is driven well into the positive region, so that grid current flows and power is consumed in the grid circuit. The smaller the operating angle, the greater the driving voltage and the larger the grid driving power required to develop full output in the load resistance. The best compromise between driving power, plate efficiency, and power output usually results when the minimum plate voltage (at the peak of the driving cycle, when the plate current reaches its highest value) is just equal to the peak positive grid voltage. Under these conditions the operating angle is usually between 150 and 180 degrees and the plate efficiency lies in the range of 70 to 80 percent. While higher plate efficiencies are possible, attaining them requires excessive driving power and grid bias, together with higher plate voltage than is "normal" for the particular tube type.

With proper design and adjustment, a Class C amplifier can be made to operate in such a way that the power input and output are proportional to the square of the applied plate voltage. This is an important consideration when the amplifier is to be plate-modulated for radiotelephony, as described in the chapter on amplitude modulation.

● FEEDBACK

It is possible to take a part of the amplified energy in the plate circuit of an amplifier and insert it into the grid circuit. When this is done the amplifier is said to have **feedback.**

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 feedback is called **negative,** or **degenerative.** On the other hand, if the voltage is fed back in phase with the grid signal, the feedback is called **positive,** or **regenerative.**

Negative Feedback

With negative feedback the voltage that is fed back opposes the signal voltage. This decreases the amplitude of the voltage acting between the grid and cathode and thus has the effect of reducing the voltage amplification. That is, a larger exciting voltage is required for obtaining the same output voltage from the plate circuit.

The greater the amount of negative feedback (when properly applied) the more independent the amplification becomes of tube characteristics and circuit conditions. This tends to make the frequency-response characteristic of the amplifier **flat** — that is, the 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." Amplifiers with negative feedback are therefore comparatively free from harmonic distortion. These advantages are worth while if the amplifier otherwise has enough voltage gain for its intended use.

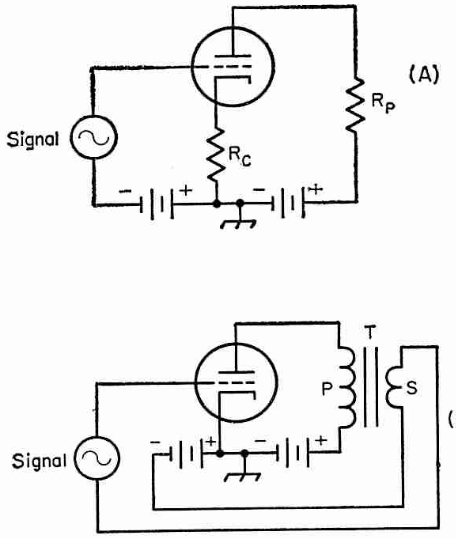


Fig. 3-14—Simple circuits for producing feedback.

In the circuit shown at A in Fig. 3-14 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. The output voltage across R_c opposes the signal voltage, so the actual a.c. voltage between the grid and cathode is equal to the *difference* between the two voltages.

The circuit shown at B in Fig. 3-14 can be used to give either negative or positive feedback. The secondary of a transformer is connected back into the grid circuit to insert a desired amount of feedback voltage. Reversing the terminals of either transformer winding (but not both simultaneously) will reverse the phase.

Positive Feedback

Positive feedback increases the amplification because the feedback voltage adds to the original signal voltage and the resulting larger voltage on the grid causes a larger output voltage. The amplification tends to be greatest at one frequency (which depends upon the particular circuit arrangement) and harmonic distortion is increased. If enough energy is fed back, a self-sustaining **oscillation** — in which energy at essentially one frequency is generated by the tube itself — will be set up. In such case all the signal voltage on the grid can be supplied from the plate circuit; no external signal is needed because 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. Positive feedback finds a major application in such "oscillators," and in addition is used for selective amplification at both audio and radio frequencies, the feedback being kept below the value that causes self-oscillation.

● INTERELECTRODE CAPACITANCES

Each pair of elements in a tube forms a small capacitor, with each element acting as a capacitor "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 having a resistive load are 180 degrees out of phase, using the cathode of the tube as a reference point. However, these two voltages are *in* phase going 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.

As a result, a capacitive current flows around the circuit, its amplitude being directly proportional to the sum of the a.c. grid and plate voltages and to the grid-plate capacitance. The source of grid signal must furnish this amount of current, in addition to the capacitive current that flows in the grid-cathode capacitance. Hence the signal source "sees" an effective capacitance that is larger than the grid-cathode capacitance. This is known as the **Miller Effect.**

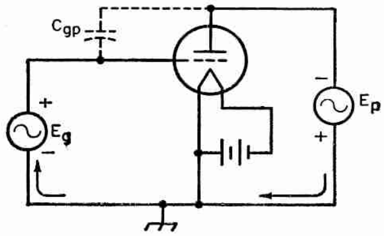


Fig. 3-15—The a.c. voltage appearing between the grid and plate of the amplifier is the sum of the signal voltage and the output voltage, as shown by this simplified circuit. Instantaneous polarities are indicated.

The greater the voltage amplification the greater the effective input capacitance. The input capacitance of a resistance-coupled amplifier is given by the formula

$$C_{\text{input}} = C_{\text{gk}} + C_{\text{gp}}(A + 1)$$

where C_{gk} is the grid-to-cathode capacitance, C_{gp} is the grid-to-plate capacitance, and A is the voltage amplification. The input capacitance may be as much as several hundred micromicrofarads when the voltage amplification is large, even though the interelectrode capacitances are quite small.

Output Capacitance

The principal component of the output capacitance of an amplifier is the actual plate-to-cathode capacitance of the tube. The output capacitance usually need not be considered in audio amplifiers, but becomes of importance at radio frequencies.

Tube Capacitance at R.F.

At radio frequencies the reactances of even very small interelectrode capacitances drop to very low values. A resistance-coupled amplifier gives very little amplification at r.f., for example, because the reactances of the interelectrode "capacitors" are so low that they practically short-circuit the input and output circuits and thus the tube is unable to amplify. This is overcome at radio frequencies by using tuned circuits for the grid and plate, making the tube capacitances part of the tuning capacitances. In this way the circuits can have the high resistive impedances necessary for satisfactory amplification.

The grid-plate capacitance is important at radio frequencies because its reactance, relatively low at r.f., offers a path over which energy can be fed back from the plate to the grid. In practically every case the feedback is in the right phase and of sufficient amplitude to cause self-oscillation, so the circuit becomes useless as an amplifier.

Special "neutralizing" circuits can be used to prevent feedback but they are, in general, not too satisfactory when used in radio receivers. They are, however, used in transmitters.

● SCREEN-GRID TUBES

The grid-plate capacitance can be 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 an electrostatic shield to prevent capacitive coupling 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.

Because of the shielding action of the screen grid, the positively charged plate cannot attract electrons from the cathode 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 then are attracted to the plate. A certain proportion do strike the screen, however, with the result that some current also flows in the screen-grid circuit.

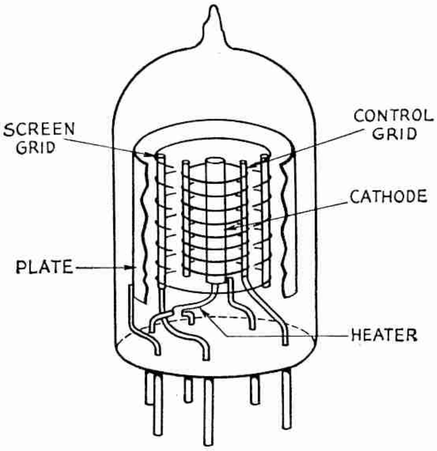


Fig. 3-16—Representative arrangement of elements in a screen-grid tetrode, with part of plate and screen cut away. This is "single-ended" construction with a button base, typical of miniature receiving tubes. To reduce capacitance between control grid and plate the leads from these elements are brought out at opposite sides; actual tubes probably would have additional shielding between these leads.

To be a good shield, the screen grid must be connected to the cathode through a circuit that has low impedance at the frequency being amplified. A bypass capacitor from screen grid to cathode, having a reactance of not more than a few hundred ohms, is generally used.

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 acts as a shield between the screen grid and plate so the secondary electrons cannot be attracted by the screen grid. They are hence attracted 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 a change in 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.

In practical screen-grid tubes the grid-plate capacitance is only a small fraction of a micromicrofarad. This capacitance is too small to cause an appreciable increase in input capacitance as described in the preceding section, so the input capacitance of a screen-grid tube is simply the sum of its grid-cathode capacitance and control-grid-to-screen capacitance. The output capacitance of a screen-grid tube is equal to the capacitance between the plate and screen.

In addition to their applications as radio-frequency amplifiers, pentodes or tetrodes also are used for audio-frequency power amplification. In tubes designed for this purpose 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 with triodes of the same power output, although harmonic distortion is somewhat greater.

Beam Tubes

A **beam tetrode** is a four-element screen-grid tube constructed in such a way that the electrons are formed into concentrated beams on their way to the plate. Additional design features overcome the effects of secondary emission so that a suppressor grid is not needed. The "beam" construction makes it possible to draw large plate currents at relatively low plate voltages, and increases the power sensitivity.

For power amplification at both audio and radio frequencies beam tetrodes have largely supplanted the non-beam types because large power outputs can be secured with very small amounts of grid driving power.

Variable-μ Tubes

The mutual conductance of a vacuum tube decreases when its grid bias is made more negative, 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.

The ordinary type of tube has what is known as a **sharp-cutoff** characteristic. The mutual conductance decreases at a uniform rate as the negative bias is increased. The amount of signal voltage that such a tube can handle without causing distortion is 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 decreases with increasing grid bias. The variable-μ tube can handle a much larger signal than the sharp-cutoff type before the signal swings either beyond the zero grid-bias point or the plate-current cutoff point.

● INPUT AND OUTPUT IMPEDANCES

The **input impedance** of a vacuum-tube amplifier is the impedance "seen" by the signal source when connected to the input terminals of the amplifier. In the types of amplifiers previously discussed, the input impedance is the impedance measured between the grid and cathode of the tube with operating voltages applied. At audio frequencies the input impedance of a Class A_1 amplifier is for all practical purposes the input capacitance of the stage. If the tube is driven into the grid-current region there is in addition a resistance component in the input impedance, the resistance having an average value equal to E^2/P, where E is the r.m.s. driving voltage and P is the power in watts consumed in the grid. The resistance usually will vary during the a.c. cycle because grid current may flow only during part of the cycle; also, the grid-voltage/grid-current characteristic is seldom linear.

The **output impedance** of amplifiers of this type consists of the plate resistance of the tube shunted by the output capacitance.

At radio frequencies, when tuned circuits are employed, the input and output impedances are usually pure resistances; any reactive components are "tuned out" in the process of adjusting the circuits to resonance at the operating frequency.

● 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 meeting point for the input and output circuits. However, it is possible to use any one of the three principal elements as the common point. This leads to two additional 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. 3-17. 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

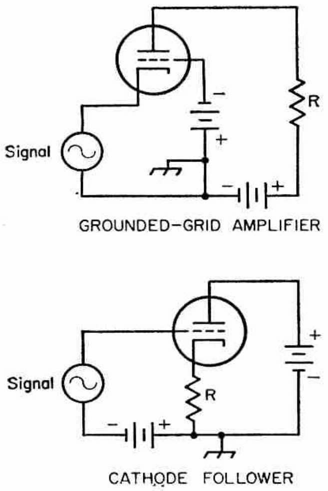


Fig. 3-17—In the upper circuit, the grid is the junction point between the input and output circuits. In the lower drawing, the plate is the junction. In either case the output is developed in the load resistor, R, and may be coupled to a following amplifier by the usual methods.

grid is thus the common element. The a.c. component of the plate current has to flow through the signal source to reach the cathode. The source of signal is in series with the load through the plate-to-cathode resistance of the tube, so some of the power in the load is supplied by the signal source. In transmitting applications this fed-through power is of the order of 10 per cent of the total power output, using tubes suitable for grounded-grid service.

The input impedance of the grounded-grid amplifier consists of a capacitance in parallel with an equivalent resistance representing the power furnished by the driving source to the grid and to the load. This resistance is of the order of a few hundred ohms. The output impedance, neglecting the interelectrode capacitances, is equal to the plate resistance of the tube. This is the same as in the case of the grounded-cathode amplifier.

The grounded-grid amplifier is widely used 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 feedback that causes oscillation. This requires that the grid act as a shield between the cathode and plate, reducing the plate-cathode capacitance to a very low value.

Cathode Follower

The cathode follower uses the plate of the tube as the common element. The input signal is applied between the grid and plate (assuming negligible impedance in the batteries) and the output is taken between cathode and plate. This circuit is degenerative; in fact, all of the output voltage is fed back into the input circuit out of phase with the grid signal. The input signal therefore has to be larger than the output voltage; that is, the cathode follower gives a loss in voltage, although it gives the same power gain as other circuits under equivalent operating conditions.

An important feature of the cathode follower is its low output impedance, which is given by the formula (neglecting interelectrode capacitances)

$$Z_{\text{out}} = \frac{r_p}{1 + \mu}$$

where r_p is the tube plate resistance and μ is the amplification factor. Low output impedance is a valuable characteristic in an amplifier designed to cover a wide band of frequencies. In addition, the input capacitance is only a fraction of the grid-to-cathode capacitance of the tube, a feature of further benefit in a wide-band amplifier. The cathode follower is useful as a step-down impedance transformer, since the input impedance is high and the output impedance is low.

● 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 usually make a rectifier-type 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.

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. 3-18. 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. The balance is never quite perfect, however, so filament-type tubes are never completely hum-

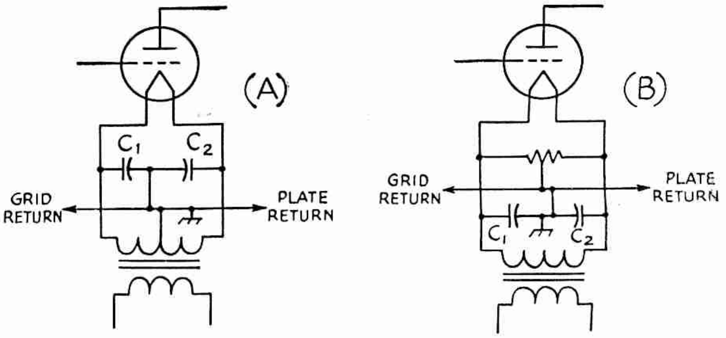


Fig. 3-18—Filament center-tapping methods for use with directly heated tubes.

free. For this reason directly-heated filaments are employed for the most part in power tubes, where the amount of hum introduced is extremely small in comparison with the power-output level.

With indirectly heated cathodes the chief problem is the magnetic field set up by the heater. Occasionally, also, there is leakage between the heater and cathode, allowing a small a.c. voltage to get to the grid. If hum appears, grounding one side of the heater supply usually will help to reduce it, although sometimes better results are obtained if the heater supply is center-tapped and the center-tap grounded, as in Fig. 3-18.

Cathode Bias

In the simplified amplifier circuits discussed in this chapter, grid bias has been supplied by a battery. However, in equipment that operates from the power line **cathode bias** is very frequently used.

The cathode-bias method uses a resistor (**cathode resistor**) connected in series with the cathode, as shown at *R* in Fig. 3-19. 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.

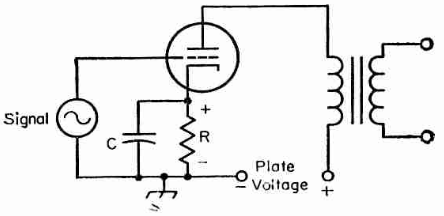


Fig. 3-19—Cathode biasing. *R* is the cathode resistor and C is the cathode bypass capacitor.

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 bypassed by a capacitor, *C*, that has very low reactance compared with the resistance of *R*. Depending on the type of tube and the particular kind of operation, *R* may be between about 100 and 3000 ohms. For good bypassing at the low audio frequencies, *C* should be 10 to 50 microfarads (electrolytic capacitors are used for this purpose). At radio frequencies, capacitances of about 100 μμf. to 0.1 μf. 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 μf. is satisfactory.

The value of cathode resistor for an amplifier having negligible d.c. resistance in its plate circuit (transformer or impedance coupled) 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.

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 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.

Calculation of the cathode resistor for a resistance-coupled amplifier is ordinarily not practicable by the method described above, because the plate current in such an amplifier is usually much smaller than the rated value given in the tube tables. However, representative data for the tubes commonly used as resistance-coupled amplifiers are given in the chapter on audio amplifiers, including cathode-resistor values.

"Contact Potential" Bias

In the absence of any negative bias voltage on the grid of a tube, some of the electrons in the space charge will have enough velocity to reach the grid. This causes a small current (of the order of microamperes) to flow in the external circuit between the grid and cathode. If the current is made to flow through a high resistance — a megohm or so — the resulting voltage drop in the resistor will give the grid a negative bias of the order of one volt. The bias so obtained is called contact-potential bias.

Contact-potential bias can be used to advantage in circuits operating at low signal levels (less than one volt peak) since it eliminates the cathode-bias resistor and bypass capacitor. It is principally used in low-level resistance-coupled audio amplifiers. The bias resistor is connected directly between grid and cathode, and must be isolated from the signal source by a blocking capacitor.

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. 3-20. Resistor R is the **screen dropping resistor,** and C is the **screen bypass capacitor.** 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.

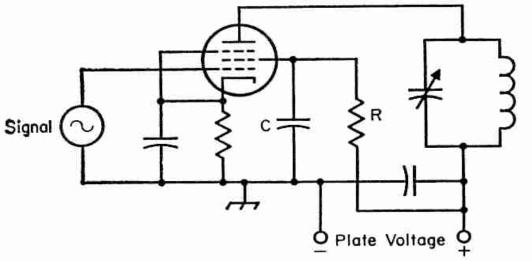


Fig. 3-20—Screen-voltage supply for a pentode tube through a dropping resistor, R. The screen bypass capacitor, C, must have low enough reactance to bring the screen to ground potential for the frequency or frequencies being amplified.

The reactance of the screen bypass capacitor, C, should be low compared with the screen-to-cathode impedance. For radio-frequency applications a capacitance in the vicinity of 0.01 μf. is amply large.

In some vacuum-tube circuits the screen voltage is obtained from a voltage divider connected across the plate supply. The design of voltage dividers is discussed at length in Chapter 7 on Power Supplies.

Oscillators

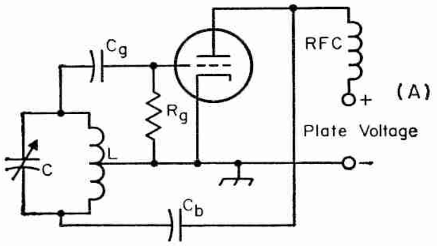
It was mentioned earlier that if there is enough positive feedback in an amplifier circuit, self-sustaining oscillations will be set up. When an amplifier is arranged so that this condition exists it is called an **oscillator.**

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. For example, in Fig. 3-21A the circuit LC is tuned to the desired frequency of oscillation. The cathode of the tube is connected to a tap on coil L and the grid and plate are connected to opposite ends of the tuned circuit. When an r.f. current flows in the tuned circuit there is a voltage drop across L that increases progressively along the turns. Thus the point at which the tap is connected will be at an intermediate potential with respect to the two ends of the coil. The amplified current in the plate circuit, which flows through the bottom section of L, is in phase with the current already flowing in the circuit and thus in the proper relationship for positive feedback.

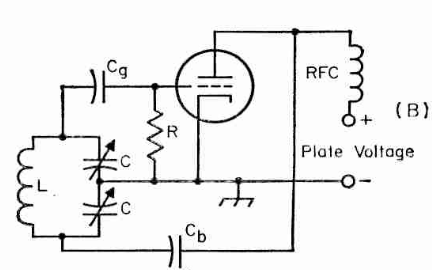
The amount of feedback depends on the position of the tap. If the tap is too near the grid end the voltage drop between grid and cathode is too small to give enough feedback to sustain oscillation, and if it is too near the plate end the impedance between the cathode and plate is too small to permit good amplification. Maximum feedback usually is obtained when the tap is somewhere near the center of the coil.

The circuit of Fig. 3-21A is parallel-fed, C_b being the blocking capacitor. The value of C_b is not critical so long as its reactance is low (not more than a few hundred ohms) at the operating frequency.

Capacitor C_g is the **grid capacitor.** It and R_g (the **grid leak**) are used for the purpose of ob-



HARTLEY CIRCUIT



COLPITTS CIRCUIT

Fig. 3-21—Basic oscillator circuits. Feedback 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 capacitor.

taining grid bias for the tube. In most oscillator circuits the tube generates its own bias. During the part of the cycle when the grid is positive with respect to the cathode, it attracts electrons. 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 (a few hundred ohms) at the operating frequency.

The circuit shown at B in Fig. 3-21 uses the voltage drops across two capacitors in series in the tuned circuit to supply the feedback. Other than this, the operation is the same as just described. The feedback 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).

Another type of oscillator, called the **tuned-plate tuned-grid** circuit, is shown in Fig. 3-22.

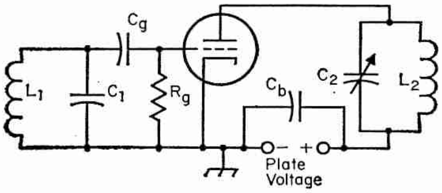


Fig. 3-22—The tuned-plate tuned-grid oscillator.

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 feedback 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 feedback 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 capacitor have the same functions as in the other circuits. In this case it is convenient to use series feed for the plate circuit, so C_b is a bypass capacitor to guide the r.f. current around the plate supply.

There are many oscillator circuits (examples of others will be found in later chapters) but the basic feature of all of them is that there is positive feedback in the proper amplitude and phase to sustain oscillation.

Oscillator Operating Characteristics

When an oscillator is delivering power to a load, the adjustment for proper feedback will depend on how heavily the oscillator is loaded — that is, how much power is being taken from the circuit. If the feedback is not large enough — **grid excitation** too small — a small increase in load may tend to throw the circuit out of oscillation. On the other hand, too much feedback will make the grid current excessively high, with the result that the power loss in the grid circuit becomes larger than necessary. Since the oscillator itself supplies this grid power, excessive feedback 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.** 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 the tuning capacitor will alter the 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.**

A change in plate voltage usually will cause the frequency to change a small amount, an effect called **dynamic instability.** Dynamic instability can be reduced by using a tuned circuit of high effective Q. The energy taken from the circuit to supply grid losses, as well as energy supplied to a load, represent an increase in the effective resistance of the tuned circuit and thus lower its Q. For highest stability, therefore, the coupling between the tuned circuit and the tube and load must be kept as loose as possible. Preferably, the oscillator should not be required to deliver power to an external circuit, and a high value of grid leak resistance should be used since this helps to raise the tube grid and plate resistances as seen by the tuned circuit. Loose coupling can be effected in a variety of ways — one, for example, is by "tapping down" on the tank for the connections to the grid and plate. This is done in the "series-tuned" Colpitts circuit widely used in variable-frequency oscillators for amateur transmitters and described in a later chapter. Alternatively, the L/C ratio may be made as small as possible while sustaining stable oscillation **(high C)** with the grid and plate connected to the ends of the circuit as shown in Figs. 3-21 and 3-22. Using relatively high plate voltage and low plate current also is desirable.

In general, dynamic stability will be at maximum when the feedback is adjusted to the least value that permits reliable oscillation. The use of a tube having a high value of transconductance is desirable, since the higher the transconductance the looser the permissible coupling to the tuned circuit and the smaller the feedback required.

Load variations act in much the same way as plate-voltage variations. A temperature change in the load may also result in drift.

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. 3-21 and 3-22 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. 3-23 shows the Hartley circuit with the plate end of the circuit grounded. 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 capacitor, C_b, across the plate supply. Direct current flows to the cathode through the lower part of the tuned-circuit coil, *L*. An advantage of such a circuit is that the frame of the tuning capacitor can be grounded.

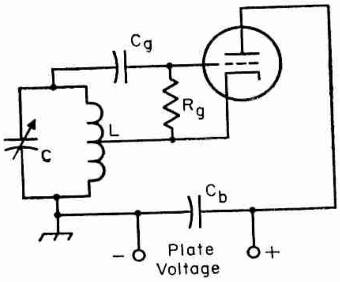


Fig. 3-23—Showing how the plate may be grounded for r.f. in a typical oscillator circuit (Hartley).

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.

Clipping Circuits

Vacuum tubes are readily adaptable to other types of operation than ordinary (without substantial distortion) amplification and the generation of single-frequency oscillations. Of particular interest is the clipper or limiter circuit, because of its several applications in receiving and other equipment.

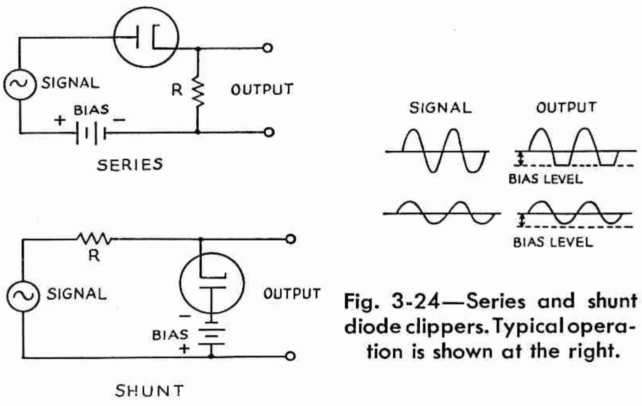


Fig. 3-24—Series and shunt diode clippers. Typical operation is shown at the right.

Diode Clipper Circuits

Basic diode clipper circuits are shown in Fig. 3-24. In the series type a positive d.c. bias voltage is applied to the plate of the diode so it is normally conducting. When a signal is applied the current through the diode will change proportionately during the time the signal voltage is positive at the diode plate and for that part of the negative half of the signal during which the instantaneous voltage does not exceed the bias. When the negative signal voltage exceeds the positive bias the resultant voltage at the diode plate is negative and there is no conduction. Thus part of the negative half cycle is clipped as shown in the drawing at the right. The level at which clipping occurs depends on the bias voltage, and the proportion of signal clipping depends on the signal strength in relation to the bias voltage. If the peak signal voltage is below the bias level there is no clipping and the output wave shape is the same as the input wave shape, as shown in the lower sketch. The output voltage results from the current flow through the load resistor *R*.

In the shunt-type diode clipper negative bias is applied to the plate so the diode is normally nonconducting. In this case the signal voltage is fed through the series resistor *R* to the output circuit (which must have high impedance compared with the resistance of *R*). When the negative half of the signal voltage exceeds the bias voltage the diode conducts, and because of the voltage drop in *R* when current flows the output voltage is reduced. By proper choice of *R* in relationship to the load on the output circuit the clipping can be made equivalent to that given by the series circuit. There is no clipping when the peak signal voltage is below the bias level.

Two diode circuits can be combined so that both the negative and positive peaks of the signal are clipped.

Triode Clippers

The circuit shown at A in Fig. 3-25 is capable of clipping both negative and positive signal peaks. On positive peaks its operation is similar to the shunt diode clipper, the clipping taking place when the positive peak of the signal voltage

Clipping Circuits

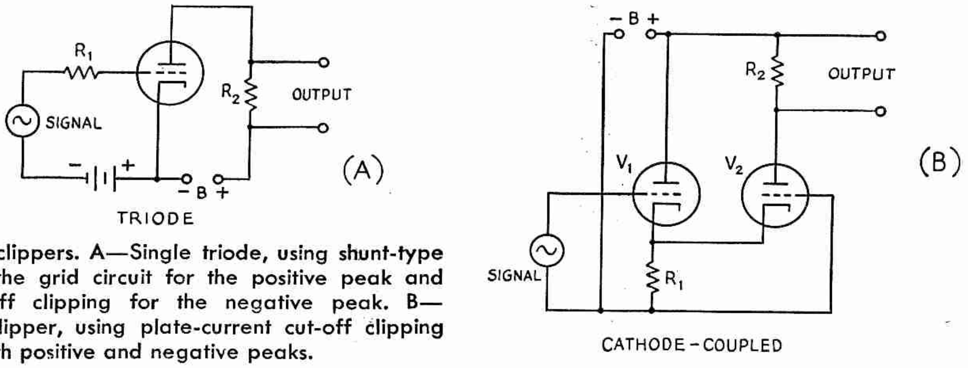


Fig. 3-25—Triode clippers. A—Single triode, using shunt-type diode clipping in the grid circuit for the positive peak and plate-current cut-off clipping for the negative peak. B—Cathode-coupled clipper, using plate-current cut-off clipping for both positive and negative peaks.

is large enough to drive the grid positive. The positive-clipped signal is amplified by the tube as a resistance-coupled amplifier. Negative peak clipping occurs when the negative peak of the signal voltage exceeds the fixed grid bias and thus cuts off the plate current in the output circuit.

In the cathode-coupled clipper shown at B in Fig. 3-25 V_1 is a cathode follower with its output circuit directly connected to the cathode of V_2, which is a grounded-grid amplifier. The tubes are biased by the voltage drop across R_1, which carries the d.c. plate currents of both tubes. When the negative peak of the signal voltage exceeds the d.c. voltage across R_1 clipping occurs in V_1, and when the positive peak exceeds the same value of voltage V_2's plate current is cut off. (The bias developed in R_1 tends to be constant because the plate current of one tube increases when the plate current of the other decreases.) Thus the circuit clips both positive and negative peaks. The clipping is symmetrical, providing the d.c. voltage drop in R_2 is small enough so that the operating conditions of the two tubes are substantially the same. For signal voltages below the clipping level the circuit operates as a normal amplifier with low distortion.

U.H.F. and Microwave Tubes

At ultrahigh frequencies, interelectrode capacitances 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, **transit time** and other 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 $\frac{1}{10}$ 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 most tubes of conventional design, the upper limit of useful operation is around 150 Mc. For higher frequencies tubes of special construction are required. About the only means available for reducing interelectrode capacitances is to reduce the physical size of the elements, which is practical only in tubes which do not have to handle appreciable power. However, it is possible to reduce the internal lead inductance very materially by minimizing the lead length and by using two or more leads in parallel from an electrode.

In some types the electrodes are provided with up to five separate leads which may be connected in parallel externally. 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. With linear tank circuits the leads become a part of the line and have distributed rather than lumped constants.

In "lighthouse" tubes or disk-seal tubes, the plate, grid and cathode are assembled in parallel planes, as shown in Fig. 3-26, instead of coaxially. The disk-seal terminals practically eliminate lead inductance.

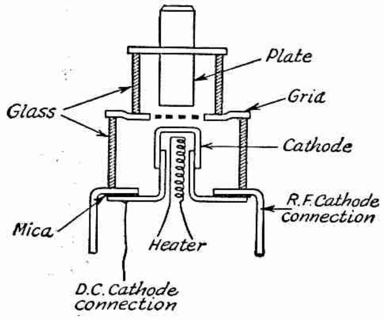


Fig. 3-26—Sectional view of the "lighthouse" tube's construction. Close electrode spacing reduces transit time while the disk electrode connections reduce lead inductance.

Velocity Modulation

In conventional tube operation the potential on the grid tends to reduce the electron velocity during the more negative half of the cycle, while on the other half cycle the positive potential on the grid serves to accelerate the electrons. 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 is turned to 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.

The velocity modulation principle may be used in a number of ways, leading to several tube designs. The major tube of this type is the "klystron."

The Klystron

In the **klystron** tube the electrons emitted by the cathode pass through an electric field established by two grids in a cavity resonator 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 slower-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 then goes to a **catcher** cavity where it again passes through two parallel grids, and the r.f. current created by the bunching of the electron beam induces an r.f. voltage between the grids. The catcher cavity is made resonant at the frequency of the velocity-modulated electron beam, so that an oscillating field is set up within it by the passage of the electron bunches through the grid aperture.

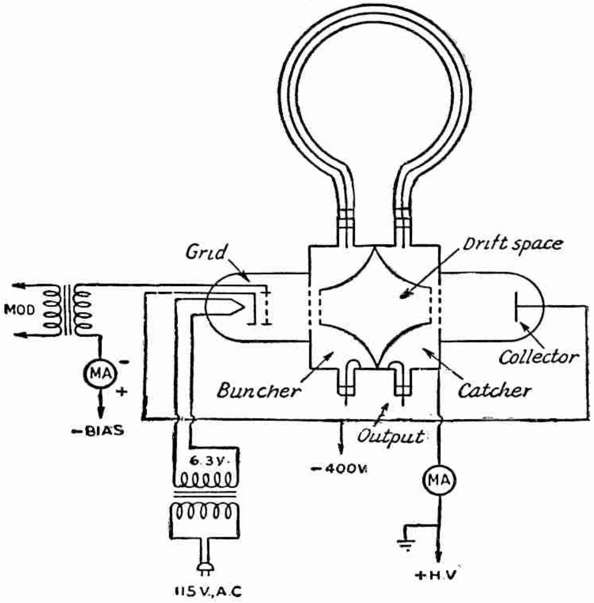


Fig. 3-27—Circuit diagram of the klystron oscillator, showing the feed-back loop coupling the frequency-controlling cavities.

If a feed back loop is provided between the two cavities, as shown in Fig. 3-27, 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 cavities. Although the bunched beam current is rich in harmonics the output wave form is remarkably pure because the high *Q* of the catcher cavity suppresses the unwanted harmonics.

Magnetrons

A **magnetron** is fundamentally a diode with cylindrical electrodes placed in a uniform magnetic field, with the lines of magnetic force parallel to the axes of the elements. The simple cylindrical magnetron consists of a cathode surrounded by a concentric cylindrical anode. In the more efficient split-anode magnetron the cylinder is divided lengthwise.

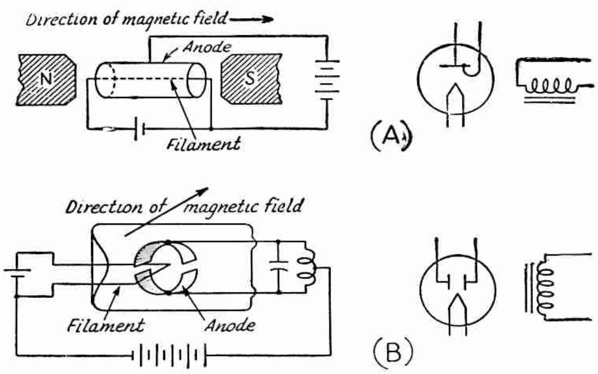


Fig. 3-28—Conventional magnetrons, with equivalent schematic symbols at the right. A, simple cylindrical magnetron. B, split-anode negative-resistance magnetron.

Magnetron oscillators are operated in two different ways. Electrically the circuits are similar, the difference being in the relation between eletron 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 the half of the anode that is at the lower potential. That is, 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.

In the transit-time magnetron the frequency is determined primarily by the tube dimensions and by the electric and magnetic field intensities rather than by the tuning of the tank circuits. 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. An alternating voltage applied between the two halves of the anode will cause the potentials of these halves to vary about their average positive values. If the period (time required for one cycle) 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 used to sustain oscillations in a resonant transmission line connected between the two halves of the anode.

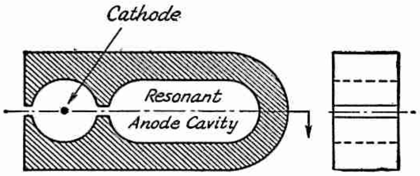


Fig. 3-29—Split-anode magnetron with integral resonant anode cavity for use at u.h.f.

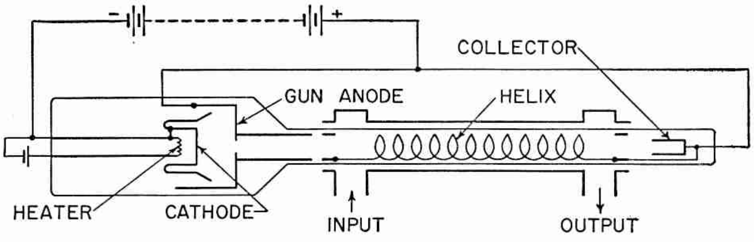


Fig. 3-30—Schematic drawing of a traveling-wave amplifier tube.

Split-anode magnetrons for u.h.f. are constructed with a cavity resonator built into the tube structure, as illustrated in Fig. 3-29. 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 4 to 16 or more segments, the resonant cavities for each anode being coupled to the common cathode region by slots of critical dimensions.

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.

Traveling-Wave Tubes

Gains as high as 23 db. over a bandwidth of 800 Mc. at a center frequency of 3600 Mc. have been obtained through the use of a **traveling-wave** amplifier tube shown schematically in Fig. 3-30. An electromagnetic wave travels down the helix, and an electron beam is shot through the helix parallel to its axis, and in the direction of propagation of the wave. When the electron velocity is about the same as the wave velocity in the absence of the electrons, turning on the electron beam causes a power gain for wave propagation in the direction of the electron motion.

The portions of Fig. 3-30 marked "input" and "output" are waveguide sections to which the ends of the helix are coupled. In practice two electromagnetic focusing coils are used, one forming a lens at the electron gun end, and the other a solenoid running the length of the helix.

The outstanding features of the traveling-wave amplifier tube are its great bandwidth and large power gain. However, the efficiency is rather low. Typical power output is of the order of 200 milliwatts.

Semiconductor Devices

Certain materials whose resistivity is not high enough to classify them as good insulators, but is still high compared with the resistivity of common metals, are known as **semiconductors.** These materials, of which germanium and silicon are examples, have an atomic structure that normally is associated with insulators. However, when small amounts of impurities are introduced during the manufacture of germanium or silicon crystals, it is possible for free electrons to exist and to move through the crystals under the influence of an electric field. It is also possible for some of the atoms to be deficient in an electron, and these electron deficiencies or **holes** can move from atom to atom when urged to do so by an applied electric force. (The movement of a hole is actually the movement of an electron, the electron becoming detached from one atom, making a hole in that atom, in order to move into an existing hole in another atom.) The holes can be considered to be equivalent to particles carrying a positive electric charge, while the electrons of course have negative charges. Holes and electrons are called charge **carriers** in semiconductors.

Electron and Hole Conduction

Material which conducts by virtue of a deficiency in electrons — that is, by **hole conduction** — is called **p-type** material. In **n-type** material, which has an excess of electrons, the conduction is termed "electronic." If a piece of p-type material is joined to a piece of n-type material as at A in Fig. 4-1 and a voltage is applied to the pair as at B, current will flow across the boundary or junction between the two (and also in the external circuit) when the battery has the polarity indicated. Electrons, indicated by the minus symbol, are attracted across the junction from the n material through the p material to the positive terminal of the battery, and holes, indicated by the plus symbol, are attracted in the opposite direction across the junction by the negative potential of the battery. Thus current flows through the circuit by means of electrons moving one way and holes the other.

If the battery polarity is reversed, as at C, the excess electrons in the n material are attracted away from the junction and the holes in the p material are attracted by the negative potential of the battery away from the junction. This leaves the junction region without any current carriers, consequently there is no conduction.

In other words, a junction of p- and n-type materials constitutes a rectifier. It differs from the tube diode rectifier in that there is a measurable, although comparatively very small, reverse current. The reverse current results from the presence of some carriers of the type opposite to those which principally characterize the material. The principal ones are called **majority carriers,** while the lesser ones are **minority carriers.**

The process by which the carriers cross the junction is essentially diffusion, and takes place comparatively slowly. This, together with the fact that the junction forms a capacitor with the two plates separated by practically zero spacing and hence has relatively high capacitance, places a limit on the upper frequency at which semiconductor devices of this construction will operate, as compared with vacuum tubes. Also, the number of excess electrons and holes in the material depends upon temperature, and since the conductivity in turn depends on the number of excess holes and electrons, the device is more temperature sensitive than is a vacuum tube.

Capacitance may be reduced by making the contact area very small. This is done by means of a **point contact,** a tiny p-type region being formed under the contact point during manufacture when n-type material is used for the main body of the device.

SEMICONDUCTOR DIODES

Diodes of the point-contact type are used for many of the same purposes for which tube diodes are used. The construction of such a diode is

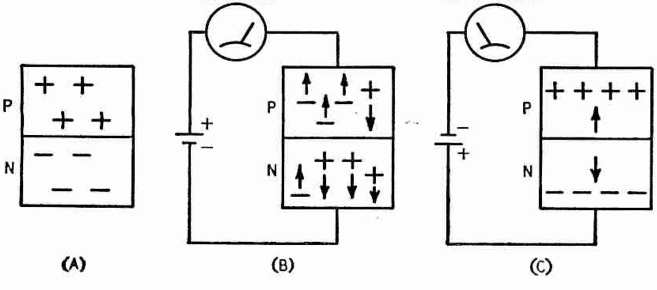


Fig. 4-1—A p-n junction (A) and its behavior when conducting (B) and non-conducting (C).

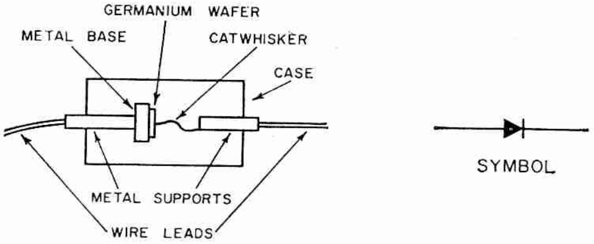


Fig. 4-2—Construction of a germanium-point-contact diode. In the circuit symbol for a contact rectifier the arrow points in the direction of minimum resistance measured by the conventional method—that is, going from the positive terminal of the voltage source through the rectifier to the negative terminal of the source. The arrow thus corresponds to the plate and the bar to the cathode of a tube diode.

shown in Fig. 4-2. Germanium and silicon are the most widely used materials, the latter principally in the u.h.f. region.

As compared with the tube diode for r.f. applications, the crystal diode has the advantages of very small size, very low interelectrode capacitance (of the order of 1 $\mu\mu$f. or less) and requires no heater or filament power.

Characteristic Curves

The germanium crystal diode is characterized by relatively large current flow with small applied voltages in the "forward" direction, and small, although finite, current flow in the reverse or "back" direction for much larger applied voltages. A typical characteristic curve is shown in Fig. 4-3. The dynamic resistance in either the forward or back direction is determined by the change in current that occurs, at any given point on the curve, when the applied voltage is changed by a small amount. The forward resistance shows some variation in the region of very small applied voltages, but the curve is for the most part quite straight, indicating fairly constant dynamic resistance. For small applied voltages, the forward resistance is of the order of 200 ohms in most such diodes. The back resistance shows considerable variation, depending on the particular voltage chosen for the measurement. It may run from a few hundred thousand ohms to over a megohm. In applications such as meter rectifiers for r.f. indicating instruments (r.f. voltmeters, wavemeter indicators, and so on) where the load resistance may be small and the applied voltage of the order of several volts, the resistances vary with the value of the applied voltage and are considerably lower.

Junction Diodes

Junction-type diodes made of germanium or silicon are employed principally as power rectifiers, in applications similar to those where selenium rectifiers are used. Depending on the design of the particular diode, they are capable of rectifying currents up to several hundred milliamperes. The safe inverse peak voltage of a junction is relatively low, so an appropriate number of rectifiers must be connected in series to operate safely on a given a.c. input voltage.

Ratings

Crystal diodes are rated primarily in terms of **maximum safe inverse voltage** and **maximum average rectified current.** Inverse voltage is a voltage applied in the direction opposite to that which causes maximum current flow. The average current is that which would be read by a d.c. meter connected in the current path.

It is also customary to specify standards of performance with respect to forward and back current. A minimum value of forward current is usually specified for one volt applied. The voltage at which the maximum tolerable back current is specified varies with the type of diode.

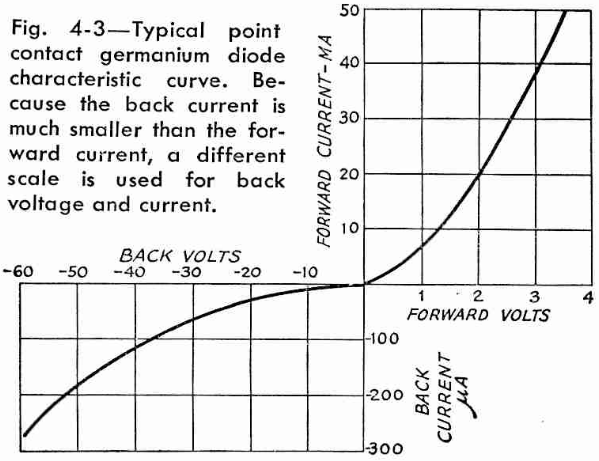


Fig. 4-3—Typical point contact germanium diode characteristic curve. Because the back current is much smaller than the forward current, a different scale is used for back voltage and current.

Zener Diodes

The "zener diode" is a special type of silicon junction diode that has a characteristic similar to that shown in Fig. 4-4. The sharp break from non-conductance to conductance is called the Zener Knee; at applied voltages greater than this breakdown point, the voltage drop across the diode is essentially constant over a wide range of currents. The substantially constant voltage

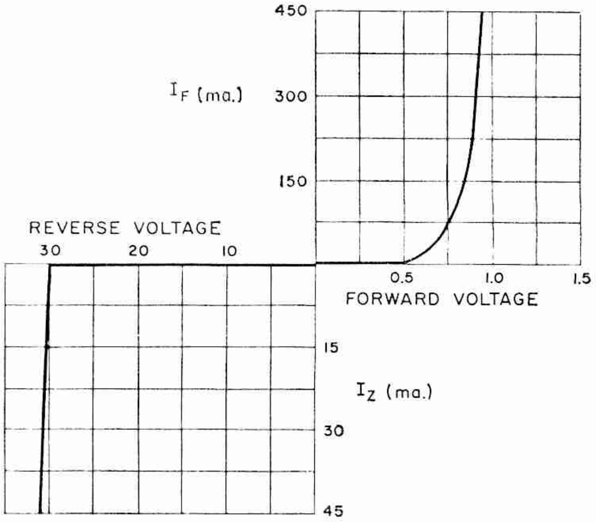


Fig. 4-4—Typical characteristic of a zener diode. In this example, the voltage drop is substantially constant at 30 volts in the (normally) reverse direction. Compare with Fig. 4-3. A diode with this characteristic would be called a "30-volt zener diode."

drop over a wide range of currents allows this semiconductor device to be used as a constant voltage reference or control element, in a manner somewhat similar to the gaseous voltage-regulator tube. Voltages for zener diode action range from a few volts to several hundred and power ratings run from a fraction of a watt to 50 watts.

Zener diodes can be connected in series to advantage; the temperature coefficient is improved over that of a single diode of equivalent rating and the power-handling capability is increased.

Two zener diodes connected in opposition, Fig. 4-5, form a simple and highly effective clipper.

Voltage-Variable Capacitors

Voltage-variable capacitors are p-n junction diodes that behave as capacitors of reasonable Q (35 or more) up to 50 Mc. and higher. They are useful in many applications because the actual capacitance value is dependent upon the d.c. bias voltage that is applied. In a typical capacitor the capacitance can be varied over a 10-to-1 range with a bias change from 0 to −100 volts. The current demand on the bias supply is on the order of a few microamperes.

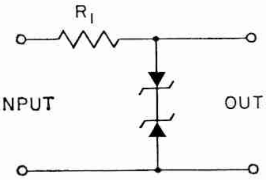


Fig. 4-5—Full-wave clipping action with two zener diodes in opposition. The output level would be at a peak-to-peak voltage of twice the zener rating of a single diode. R_1 should have a resistance value sufficient to limit the current to the zener diode rating.

Typical applications include remote control of tuned circuits, automatic frequency control of receiver local oscillators, and simple frequency modulators for communications and for sweep-tuning applications.

Transistors

Fig. 4-6 shows a "sandwich" made from two layers of p-type semiconductor material with a thin layer of n-type between. There are in effect two p-n junction diodes back to back. If a positive bias is applied to the p-type material at the left, current will flow through the left-hand junction, the holes moving to the right and the electrons from the n-type material moving to the left. Some of the holes moving into the n-type material will combine with the electrons there and be neutralized, but some of them also will travel to the region of the right-hand junction.

If the p-n combination at the right is biased negatively, as shown, there would normally be no current flow in this circuit (see Fig. 4-1C). However, there are now additional holes available at the junction to travel to point B and electrons can travel toward point A, so a current can flow even though this section of the sandwich considered alone is biased to prevent conduction. Most of the current is between A and B and does not flow out through the common connection to the n-type material in the sandwich.

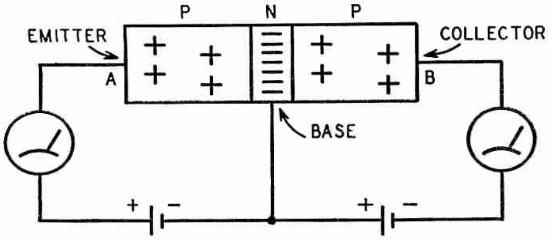


Fig. 4-6—The basic arrangement of a transistor. This represents a junction-type p-n-p unit.

A semiconductor combination of this type is called a **transistor,** and the three sections are known as the **emitter, base** and **collector,** respectively. The amplitude of the collector current depends principally upon the amplitude of the emitter current; that is, the collector current is controlled by the emitter current.

Power Amplification

Because the collector is biased in the back direction the collector-to-base resistance is high. On the other hand, the emitter and collector currents are substantially equal, so the power in the collector circuit is larger than the power in the emitter circuit ($P = I^2 R$, so the powers are proportional to the respective resistances, if the current is the same). In practical transistors emitter resistance is of the order of a few hundred ohms while the collector resistance is hundreds or thousands of times higher, so power gains of 20 to 40 db. or even more are possible.

Types

The transistor may be either of the **point-contact** or **junction** type, as shown in Fig. 4-7. Also, the assembly of p- and n-type materials may be reversed; that is, n-type material may be used instead of p-type for the emitter and collector, and p-type instead of n-type for the base. The type shown in Fig. 4-6 is a **p-n-p** transistor, while the opposite is the **n-p-n.**

Point-Contact Transistors

The point-contact transistor, shown at the left in Fig. 4-7, has two "cat whiskers" placed very close together on the surface of a germanium wafer, usually n-type material. Small p-type areas are formed under each point during manufacture. This type of construction results in quite low interelectrode capacitances, with the result that some point-contact transistors have been used at frequencies up to the v.h.f. region.

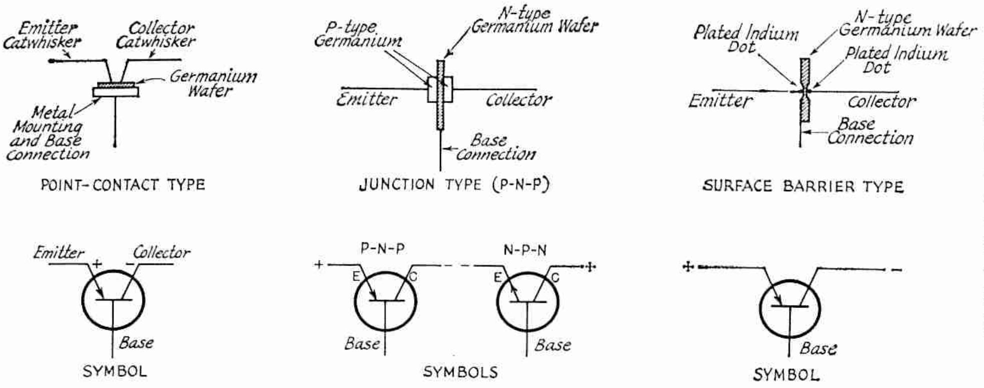


Fig. 4-7—Point-contact, junction-type and surface-barrier types of transistors with their circuit symbols. The plus and minus signs associated with the symbols indicate polarities of voltages, with respect to the base, to be applied to the elements.

The point-contact transistor is principally of historical interest, since it is now superseded by the junction type. It is difficult to manufacture, since the two contact points must be extremely close together if good characteristics are to be secured, particularly for high-frequency work.

Junction Transistors

The junction transistor, the essential construction of which is shown at the center in Fig. 4-7, has higher capacitances and higher power-handling capacity than the point-contact type. The "electrode" areas and thickness of the intermediate layer have an important effect on the upper frequency limit. Ordinary junction transistors may have cut-off frequencies (see next section) up to 20 Mc. or so. The types used for audio and low radio frequencies usually have cut-off frequencies ranging from 500 to 1000 kc.

The upper frequency limit is extended considerably in the **drift transistor.** This type has a particular form of distribution of impurities in the base material resulting in the creation of an internal electric field that accelerates the carriers across the junction. Typical drift transistors have cut-off frequencies of the order of 100 Mc.

Another type of transistor useful in high-frequency work is the **surface barrier transistor,** using plated emitter and collector electrodes on a wafer of n-type material, as shown at the right in Fig. 4-7 above. Surface barrier transistors will operate at frequencies up to 60 or 75 Mc. as amplifiers and oscillators.

● TRANSISTOR CHARACTERISTICS

An important characteristic of a transistor is its **current amplification factor,** usually designated by the symbol α. This is the ratio of the change in collector current to a small change in emitter current, measured in the common-base circuit described later, and is comparable with the voltage amplification factor (μ) of a vacuum tube. The current amplification factor is almost, but not quite, 1 in a junction transistor. It is larger than 1 in the point-contact type, values in the neighborhood of 2 being typical.

The α **cut-off frequency** is the frequency at which the current amplification drops 3 db. below its low-frequency value. Cut-off frequencies range from 500 kc. to frequencies in the v.h.f. region. The cut-off frequency indicates in a general way the frequency spread over which the transistor is useful.

Each of the three elements in the transistor has a resistance associated with it. The emitter and collector resistances were discussed earlier. There is also a certain amount of resistance associated with the base, a value of a few hundred to 1000 ohms being typical of the base resistance.

The values of all three resistances vary with the type of transistor and the operating voltages. The collector resistance, in particular, is sensitive to operating conditions.

Characteristic Curves

The operating characteristics of transistors can be shown by a series of characteristic curves. One such set of curves is shown in Fig. 4-8. It shows the collector current *vs.* collector voltage for a number of fixed values of emitter current. Practically, the collector current depends almost entirely on the emitter current and is independent of the collector voltage. The separation between curves representing equal steps of emitter current is quite uniform, indicating that almost distortionless output can be obtained over the useful operating range of the transistor.

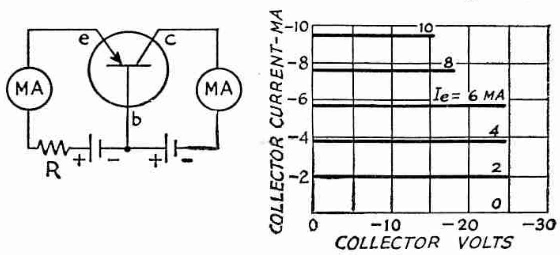


Fig. 4-8—A typical collector-current *vs.* collector-voltage characteristic of a junction-type transistor, for various emitter-current values. The circuit shows the setup for taking such measurements. Since the emitter resistance is low, a current-limiting resistor, *R*, is connected in series with the source of current. The emitter current can be set at a desired value by adjustment of this resistance.

Another type of curve is shown in Fig. 4-9, together with the circuit used for obtaining it. This also shows collector current *vs.* collector voltage, but for a number of different values of base current. In this case the emitter element is used as the common point in the circuit. The collector current is not independent of collector voltage with this type of connection, indicating that the output resistance of the device is fairly low. The base current also is quite low, which means that the resistance of the base-emitter circuit is moderately high with this method of connection. This may be contrasted with the high values of emitter current shown in Fig. 4-8.

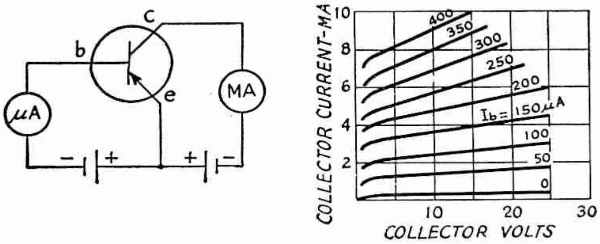


Fig. 4-9—Collector current vs. collector voltage for various values of base current, for a junction-type transistor. The values are determined by means of the circuit shown.

Ratings

The principal ratings applied to transistors are maximum collector dissipation, maximum collector voltage, maximum collector current, and maximum emitter current. The voltage and current ratings are self-explanatory.

The collector dissipation is the power, usually expressed in milliwatts, that can safely be dissipated by the transistor as heat. With some types of transistors provision is made for transferring heat rapidly through the container, and such units usually require installation on a heat "sink," or mounting that can absorb heat.

The amount of undistorted output power that can be obtained depends on the collector voltage, the collector current being practically independent of the voltage in a given transistor. Increasing the collector voltage extends the range of linear operation, but must not be carried beyond the point where either the voltage or dissipation ratings are exceeded.

● TRANSISTOR AMPLIFIERS

Amplifier circuits used with transistors fall into one of three types, known as the **grounded-base, grounded-emitter,** and **grounded-collector** circuits. These are shown in Fig. 4-10 in elementary form. The three circuits correspond approximately to the grounded-grid, grounded-cathode and cathode-follower circuits, respectively, used with vacuum tubes.

The important transistor **parameters** in these circuits are the **short-circuit current transfer ratio,** the **cut-off frequency,** and the **input** and **output impedances.** The short-circuit current transfer ratio is the ratio of a small change in output current to the change in input current that causes it, the output circuit being short-circuited. The cut-off frequency is the frequency at which the amplification decreases by 3 db. from its value at some frequency well below that at which frequency effects begin to assume importance. The input and output impedances are, respectively, the impedance which a signal source working into the transistor would see, and the internal output impedance of the transistor (corresponding to the plate resistance of a vacuum tube, for example).

Grounded-Base Circuit

The input circuit of a grounded-base amplifier must be designed for low impedance, since the emitter-to-base resistance is of the order of $25/I_e$ ohms, where I_e is the emitter current in milliamperes. The optimum output load impedance, R_L, may range from a few thousand ohms to 100,000, depending upon the requirements.

The current transfer ratio is α and the cut-off frequency is as defined previously.

In this circuit the phase of the output (collector) current is the same as that of the input (emitter) current. The parts of these currents that flow through the base resistance are likewise in phase, so the circuit tends to be regenerative and will oscillate if the current amplification factor is greater than 1. A junction transistor is stable in this circuit since α is less than 1, but a point-contact transistor will oscillate.

Grounded-Emitter Circuit

The grounded-emitter circuit shown in Fig. 4-10 corresponds to the ordinary grounded-cathode vacuum-tube amplifier. As indicated by the curves of Fig. 4-9, the base current is small and the input impedance is therefore fairly high—several thousand ohms in the average case. The collector resistance is some tens of thousands of ohms, depending on the signal source impedance. The current transfer ratio in the common-emitter circuit is equal to

$$\frac{\alpha}{1 - \alpha}$$

Since α is close to 1 (0.98 or higher being representative), the short-circuit current gain in the grounded-emitter circuit may be 50 or more. The cut-off frequency is equal to the α cut-off frequency multiplied by $(1 - \alpha)$, and therefore is relatively low. (For example, a transistor with an α cut-off of 1000 kc. and $\alpha = 0.98$ would have a cut-off frequency of $1000 \times 0.02 = 20$ kc. in the grounded-emitter circuit.)

Within its frequency limitations, the grounded-emitter circuit gives the highest power gain of the three.

In this circuit the phase of the output (collector) current is opposite to that of the input (base) current so such feedback as occurs through the small emitter resistance is negative and the amplifier is stable with either junction or point-contact transistors.

Grounded-Collector Circuit

Like the vacuum-tube cathode follower, the grounded-collector transistor amplifier has high input impedance and low output impedance. The latter is approximately equal to the impedance of the signal input source multiplied by $(1 - \alpha)$. The input resistance depends on the load resistance, being approximately equal to the load resistance divided by $(1 - \alpha)$. The fact that input resistance is directly related to the load resistance is a disadvantage of this type of amplifier if the load is one whose resistance or impedance varies with frequency.

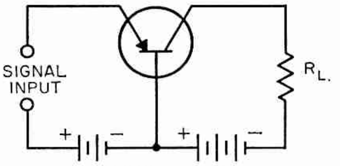
The current transfer ratio with this circuit is

$$\frac{1}{1-\alpha}$$

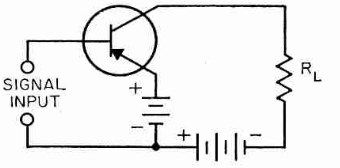
and the cut-off frequency is the same as in the grounded-emitter circuit. The output and input currents are in phase.

Practical Circuit Details

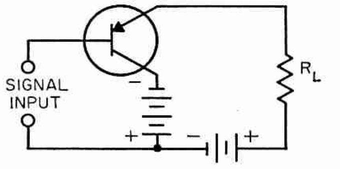
The transistor is essentially a low-voltage device, so the use of a battery power supply rather than a rectified-a.c. supply is quite common. Usually, it is more convenient to employ a single battery as a power source in preference to the two-battery arrangements shown in Fig. 4-10, so most circuits are designed for single-battery operation. Provision must be included, therefore, for obtaining proper biasing voltage for the emitter-base circuit from the battery that supplies the power in the collector circuit.



COMMON BASE



COMMON EMITTER

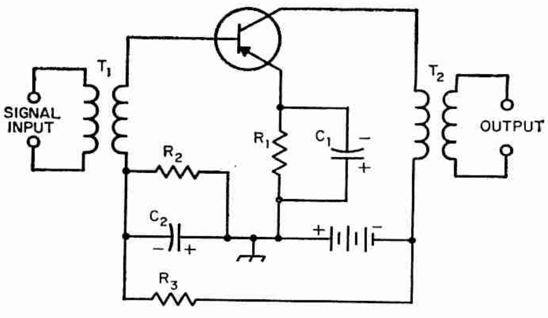


COMMON COLLECTOR

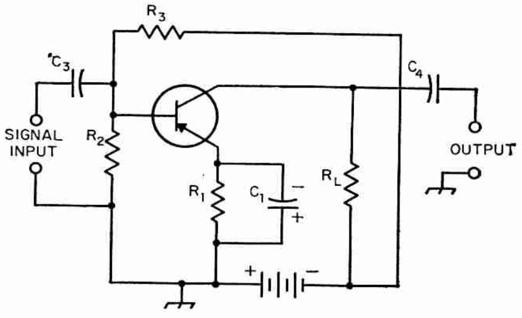
Fig. 4-10—Basic transistor amplifier circuits. R_L, the load resistance, may be an actual resistor or the primary of a transformer. The input signal may be supplied from a transformer secondary or by resistance-capacitance coupling. In any case it is to be understood that a d.c. path must exist between the base and emitter.

P-n-p transistors are shown in these circuits. If n-p-n types are used the battery polarities must be reversed.

Coupling arrangements for introducing the input signal into the circuit and for taking out the amplified signal are similar to those used with vacuum tubes. However, the actual component values will in general be quite different from those used with tubes. This is because the impedances associated with the input and output circuits of transistors may differ widely from the comparable impedances in tube circuits. Also, d.c. voltage drops in resistances may require more careful attention with transistors because of the much lower voltage available from the ordinary battery power source. Battery economy becomes an important factor in circuit design, both with respect to voltage required and to overall current drain. A bias voltage divider, for example, easily may use more power than the transistor with which it is associated.



TRANSFORMER COUPLING



RESISTANCE COUPLING

Fig. 4-11—Practical grounded-emitter circuits using transformer and resistance coupling. A combination of either also can be used—e.g., resistance-coupled input and transformer-coupled output. Tuned transformers may be used for r.f. and i.f. circuits.

With small transistors used for low-level amplification the input impedance will be of the order of 1000 ohms and the input circuit should be designed for an impedance step-down, if necessary. This can be done by appropriate choice of turns ratio for T_1 or, in the case of tuned circuits, by tapping the base down on the tuned secondary circuit. In the resistance-coupled circuit R_2 should be large compared with the input impedance, values of the order of 10,000 ohms being used.

In low-level circuits R_1 will be of the order of 1000 ohms. R_3 should be chosen to bias the transistor to the desired no-signal collector current; its value depends on R_1 and R_2 (see text).

Typical single-battery grounded-emitter circuits are shown in Fig. 4-11. R_1, in series with the emitter, is for the purpose of "swamping out" the resistance of the emitter-base diode; this swamping helps to stabilize the emitter current. The resistance of R_1 should be large compared with that of the emitter-base diode, which, as stated earlier, is approximately equal to 25 divided by the emitter current in ma.

Since the current in R_1 flows in such a direction as to bias the emitter negatively with respect to the base (a p-n-p transistor is assumed), a base-emitter bias slightly greater than the drop in R_1 must be supplied. The proper operating point is achieved through adjustment of voltage divider R_2R_3, the constants of which are chosen to give the desired value of collector current at the no-signal operating point.

In the transformer-coupled circuit, input signal currents flow through R_1 and R_2, and there would be a loss of signal power at the base-emitter diode if these resistors were not bypassed by C_1 and C_2. The capacitors should have low reactance compared with the resistances across which they are connected. In the resistance-coupled circuit R_2

has the dual function of acting as part of the bias voltage divider and as part of the load resistance for the signal-input source. Also, as seen by the signal source, R_3 is in parallel with R_2 and thus becomes part of the input load resistance. C_3 must therefore have low reactance compared with the net resistance of the parallel combination of R_2, R_3 and the base-to-emitter resistance of the transistor. The reactance of C_4 will depend on the impedance of the load into which the circuit delivers output.

The output load resistance in the transformer-coupled case will be the actual load as reflected at the primary of the transformer, and its proper value will be determined by the transistor characteristics and the type of operation (Class A, B, etc.). The value of R_L in the resistance-coupled case is usually such as to permit the maximum a.c. voltage swing in the collector circuit without undue distortion, since Class A operation is usual with this type of amplifier.

Bias Stabilization

Transistor currents are rather sensitive to temperature variations, and so the operating point tends to shift as the transistor heats. The shift in operating point unfortunately is in such a direction as to increase the heating, leading to "thermal runaway" and possible destruction of the transistor. The heat developed depends on the amount of power dissipated in the transistor, so it is obviously advantageous in this respect to operate with as little internal dissipation as possible: i.e., the d.c. input should be kept to the lowest value that will permit the type of operation desired, and in any event should never exceed the rated value for the particular transistor used.

A contributing factor to the shift in operating point is the collector-to-base leakage current (usually designated I_{co}) — that is, the current that flows from collector to base with the emitter connection open. This current, which is highly temperature sensitive, has the effect of increasing the emitter current by an amount much larger than I_{co} itself, thus shifting the operating point in such a way as to increase the collector current. This effect is reduced to the extent that I_{co} can be made to flow out of the base terminal rather than through the base-emitter diode. In the circuits of Fig. 4-11, bias stabilization is improved by making the resistance of R_1 as large as possible and both R_2 and R_3 as small as possible, consistent with other considerations such as gain and battery economy.

● TRANSISTOR OSCILLATORS

Since more power is available from the output circuit than is necessary for its generation in the input circuit, it is possible to use some of the output power to supply the input circuit and thus sustain self-oscillation. Representative oscillator circuits are shown in Fig. 4-12. Their resemblance to the similarly-named vacuum-tube circuits is evident.

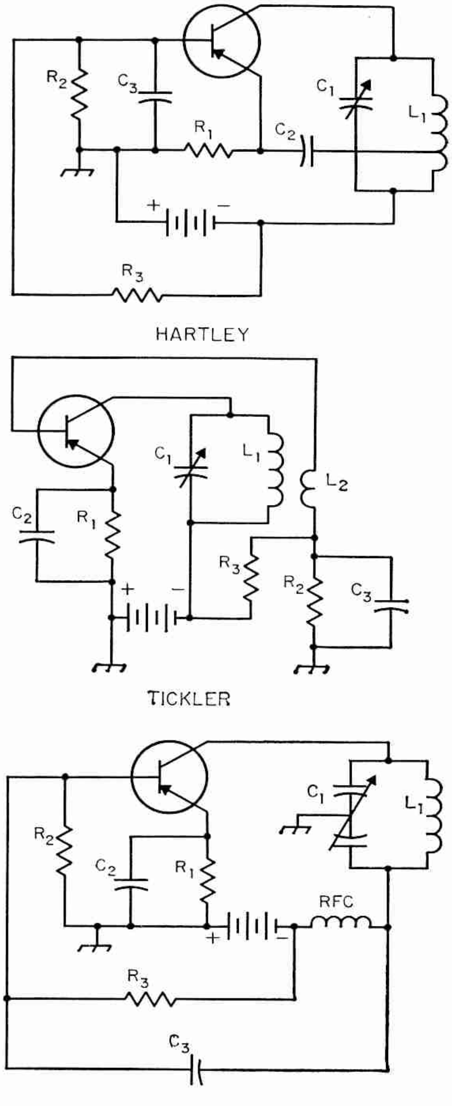


Fig. 4-12—Typical transistor oscillator circuits. Component values are discussed in the text.

The upper frequency limit for oscillation is principally a function of the cut-off frequency of the transistor used, and oscillation will cease at the frequency at which there is insufficient amplification to supply the energy required to overcome circuit losses. Transistor oscillators usually will operate up to, and sometimes well beyond, the α cut-off frequency of the particular transistor used.

The approximate oscillation frequency is that of the tuned circuit, L_1C_1. R_1, R_2 and R_3 have the same functions as in the amplifier circuits given in Fig. 4-11. Capacitors C_2 and C_3 are bypass or blocking capacitors and should have low reactance compared with the resistances with which they are associated.

Feedback in these circuits is adjusted in the same way as with tube oscillators. In the Hartley circuit it is dependent on the position of the tap on the tank coil; in the tickler circuit, on the number of turns in L_2 and degree of coupling between L_1 and L_2; and in the Colpitts circuit, on the ratio of the tank capacitance between base and emitter to the tank capacitance between collector and emitter.

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 overemphasized. In the less crowded v.h.f. bands, **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). 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, 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 major differences between receivers for phone reception and for code reception. An a.m. phone signal has side bands 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 bandwidth is less than half of this. A code signal occupies only a few hundred cycles at the most, and consequently the bandwidth of a code receiver can be small. A single-sideband phone signal takes up 3 to 4 kc., and the audio quality can be impaired if the bandwidth is much less than 3 kc. although the intelligibility will hold up down to around 2 kc. In any case, if the bandwidth of the receiver is more than necessary, signals adjacent to the desired one can be heard, and the selectivity of the receiver is less than maximum. The detection process delivers directly the audio frequencies present as modulation on an a.m. phone signal. There is no modulation on a code signal, and 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 made on the order of 500 to 1000 cycles, since these tones are within the range of optimum response of both the ear and the headset. There is no carrier frequency present in an s.s.b. signal, and this frequency must be furnished at the receiver before the audio can be recovered. The same source that is used in code reception can be utilized for the purpose. If the source of the locally generated radio frequency is a separate oscillator, the system is known as **heterodyne** reception; if the detector is made to oscillate and produce the frequency, it is known as an **autodyne** detector. Modern superheterodyne receivers generally use a separate oscillator (**beat oscillator**) to supply the locally generated frequency. Summing up the differences, phone receivers can't use as much selectivity as code receivers, and code and s.s.b. receivers require some kind of locally generated frequency to give a readable signal. Broadcast receivers can receive only a.m. phone signals because no beat oscillator is included. **Communications receivers** include beat oscillators and often some means for varying the selectivity. With high selectivity they often have a slow tuning rate.

Receiver Characteristics

Sensitivity

In commercial circles "sensitivity" 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 satisfactory definition for broadcast and communications receivers operating below about 20 Mc., where atmospheric and man-made electrical noises normally mask any noise generated by the receiver itself.

Another commercial measure of sensitivity defines it as the signal at the input of the receiver required to give a signal-plus-noise output some stated ratio (generally 10 db.) above the noise output of the receiver. This is a more useful sensitivity measure for the amateur, since it indicates how well a weak signal will be heard and is not merely a measure of the over-all amplification of the receiver. However, it is not an absolute method, because the bandwidth 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. Thermal-agitation noise is independent of frequency and is proportional to the (absolute) temperature, the resistance component of the impedance across which the thermal agitation is produced, and the bandwidth. Noise is generated in vacuum tubes by random irregularities in the current flow within them; it is convenient to express this **shot-effect noise** as an equivalent resistance in the grid circuit of a noise-free tube. This **equivalent noise resistance** is the resistance (at room temperature) that placed in the grid circuit of a noise-free tube will produce plate-circuit noise equal to that of the actual tube. The equivalent noise resistance of a vacuum tube increases with frequency.

An ideal receiver would generate no noise in its tubes and circuits, and the minimum detectable signal would be limited only by the thermal noise in the antenna. In a practical receiver, the limit is determined by how well the amplified antenna noise overrides the other noise in the plate circuit of the input stage. (It is assumed that the first stage in any good receiver will be the determining factor; the noise contributions of subsequent stages should be insignificant by comparison.) At frequencies below 20 or 30 Mc. the site noise (atmospheric and man-made noise) is generally the limiting factor.

The degree to which a practical receiver approaches the quiet ideal receiver of the same bandwidth is given by the **noise figure** of the receiver. Noise figure is defined as the ratio of the signal-to-noise power ratio of the ideal receiver to the signal-to-noise power ratio of the actual receiver output. Since the noise figure is a ratio, it is usually given in decibels; it runs around 5 to 10 db. for a good communications receiver below 30 Mc. Although noise figures of 2 to 4 db. can be obtained, they are of little or no use below 30 Mc. except in extremely quiet locations or when a very small antenna is used. The noise figure of a receiver is not modified by changes in bandwidth.

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 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 ("6 db. down" and "20 db. down").

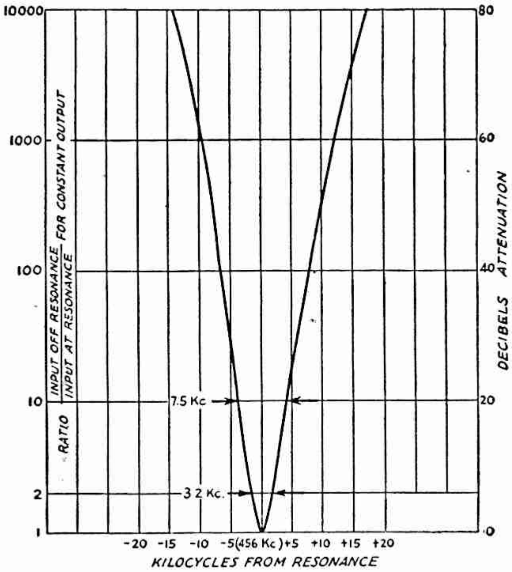


Fig. 5-1—Typical selectivity curve of a modern superheterodyne receiver. Relative response is plotted against deviations above and below the resonance frequency. The scale at the left is in terms of voltage ratios, the corresponding decibel steps are shown at the right.

The bandwidth at 6 db. down must be sufficient to pass the signal and its sidebands if faithful reproduction of the signal is desired. However, in the crowded amateur bands, it is generally advisable to sacrifice fidelity for intelligibility. The ability to reject adjacent-channel signals depends upon the **skirt selectivity** of the receiver, which is determined by the bandwidth at high attenuation. In a receiver with good skirt selectivity, the ratio of the 6-db. bandwidth to the 60-db. bandwidth will be about 0.25 for code and 0.5 for phone. The minimum usable bandwidth at 6 db. down is about 150 cycles for code reception and about 2000 cycles for phone.

Stability

The stability of a receiver is its ability to "stay put" on a signal under varying conditions of gain-control setting, temperature, supply-voltage changes and mechanical shock and distortion. 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.

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.

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 to accept signals of a specified amplitude without overloading or distortion.

Detection and Detectors

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 capacitor, C_2. The flow of rectified r.f. current causes a d.c. voltage to develop across the terminals of R_1. 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.

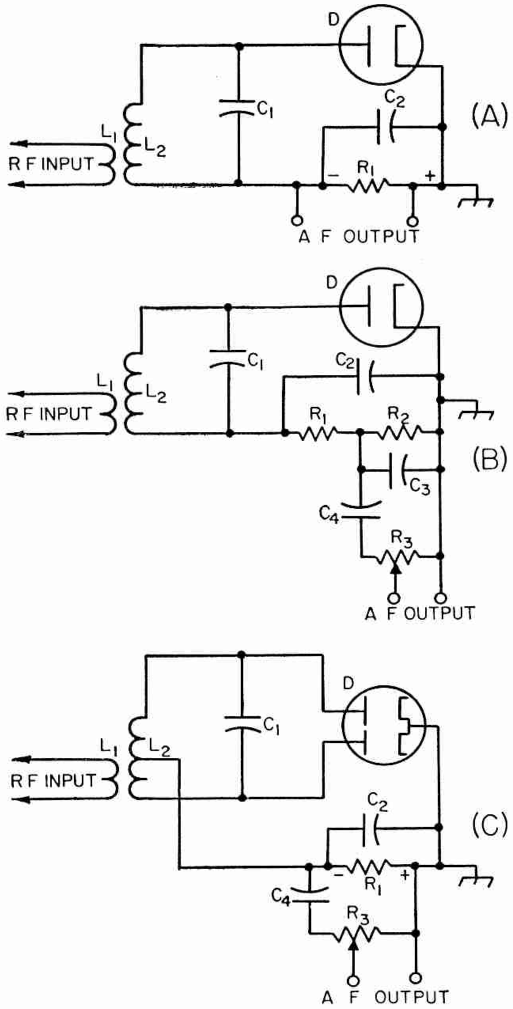


Fig. 5-2—Simplified and practical diode detector circuits. A, the elementary half-wave diode detector; B, a practical circuit, with r.f. filtering and audio output coupling; C, full-wave diode detector, with output coupling indicated. The circuit, L_2C_1, is tuned to the signal frequency; typical values for C_2 and R_1 in A and C are 250 $\mu\mu$f. and 250,000 ohms, respectively; in B, C_2 and C_3 are 100 $\mu\mu$f. each; R_1, 50,000 ohms; and R_2, 250,000 ohms. C_4 is 0.1 μf. and R_3 may be 0.5 to 1 megohm.

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. These current pulses flow in the load circuit comprised of R_1 and C_2, the resistance of R_1 and the capacity of C_2 being so proportioned that C_2 charges to the peak value of the rectified voltage on each pulse and retains enough charge between pulses so that the voltage across R_1 is smoothed out, as shown in C. C_2 thus acts as a filter for the radio-frequency component of the output of the rectifier, leaving a d.c. component that varies in the same way as the modulation on the original signal. When this varying d.c. voltage is applied to a following amplifier through a coupling capacitor (C_4 in Fig. 5-2), only the *variations* in voltage are transferred, so that the final output signal is a.c., as shown in D.

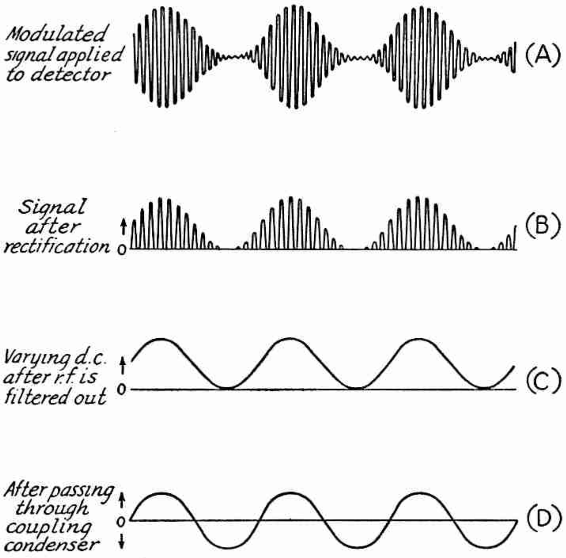


Fig. 5-3—Diagrams showing the detection process.

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 capacitor, C_4, to a load resistor, R_3, which usually is a "potentiometer" so that the audio volume can be adjusted to a desired level.

Coupling to the potentiometer (volume control) through a capacitor also avoids any flow of d.c. through the control. The flow of d.c. through a high-resistance volume 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 r.f. filtering is easier than in the half-wave circuit. As a result, less attenuation of the higher audio frequencies will be obtained for any given degree of r.f. filtering.

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. If the capacity of C_2 is too large, response at the higher audio frequencies will be lowered.

Compared with other detectors, the sensitivity of the diode is low, 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 application of a signal to the grid circuit causes an increase in average plate current. The average plate current follows the changes in signal in a fashion similar to the rectified current in a diode detector.

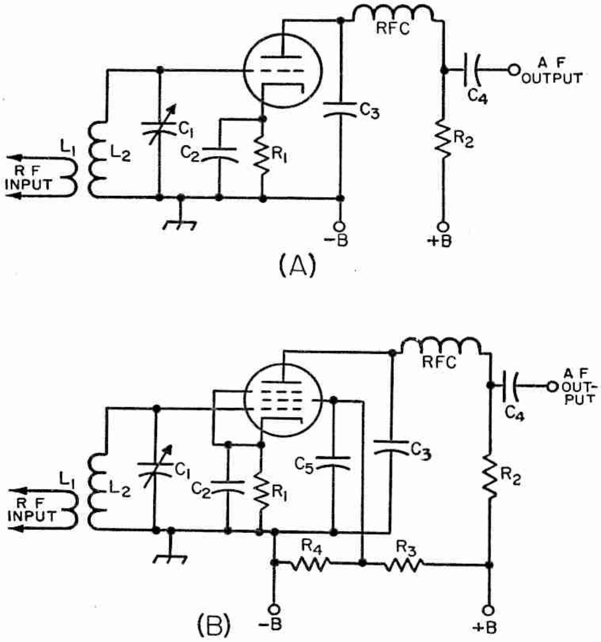


Fig. 5-4—Circuits for plate detection. A, triode; B, pentode. The input circuit, L_2C_1, is tuned to the signal frequency. Typical values for the other components are:

Component	Circuit A	Circuit B
C_2	0.5 μf. or larger.	0.5 μf. or larger.
C_3	0.001 to 0.002 μf.	250 to 500 μμf.
C_4	0.1 μf.	0.1 μf.
C_5		0.5 μf. 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.

Circuits for triodes and pentodes are given in Fig. 5-4. C_3 is the plate bypass capacitor, and, with *RFC*, prevents r.f. from appearing in the output. The cathode resistor, R_1, provides the operating grid bias, and C_2 is a bypass for both radio and audio frequencies. R_2 is the plate load resistance and C_4 is the output coupling capacitor. In the pentode circuit at B, R_3 and R_4 form a voltage divider to supply the proper screen potential (about 30 volts), and C_5 is a bypass capacitor. C_2 and 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 of any tube is very high when the bias is 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 because there is some amplifying action in the tube. It will handle large signals, but is not 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 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 feedback for the audio frequencies. The cathode resistor is bypassed for r.f. but not for audio, while the plate circuit is bypassed to

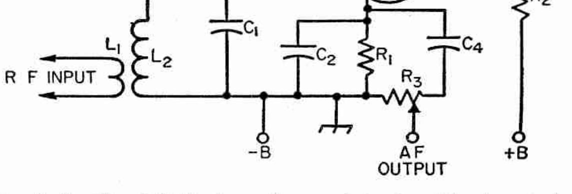


Fig. 5-5—The infinite-impedance detector. The input circuit, L_2C_1, is tuned to the signal frequency. Typical values for the other components are:

C_2—250 μμf.
C_3—0.5 μf.
C_4—0.1 μf.
R_1—0.15 megohm.
R_2—25,000 ohms.
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.

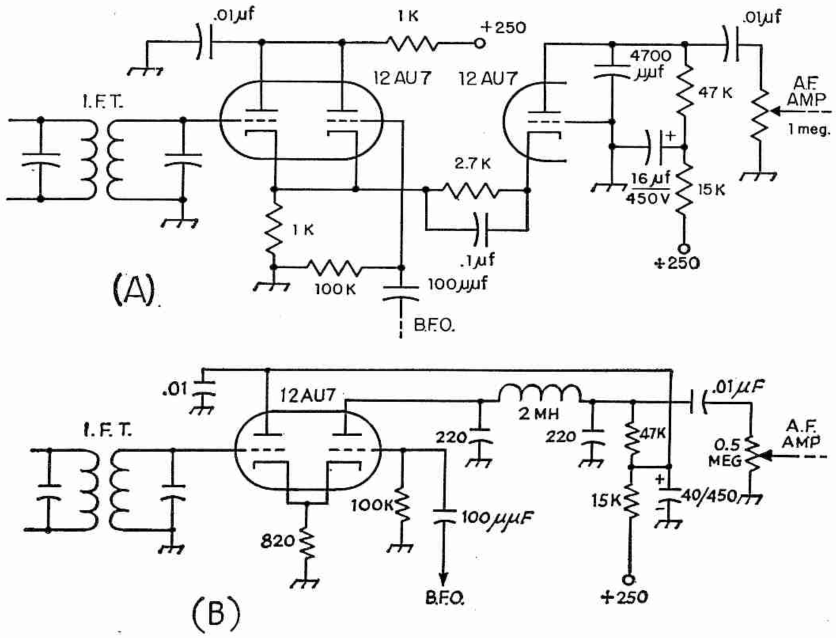


Fig. 5-6—Two versions of the "product detector" circuit. In the circuit at A separate tubes are used for the signal circuit cathode follower, the b.f.o. cathode follower and the mixer tube. In B the mixer and b.f.o. follower are combined in one tube, and a low-pass filter is used in the output.

ground for both audio and radio frequencies. An r.f. filter 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 consequently increases with signal. Because of this and the large initial drop across R_1, the grid usually cannot be driven positive by the signal, and no grid current can be drawn.

Product Detector

The **product detector** circuits of Fig. 5-6 are useful in s.s.b. and code reception because they minimize intermodulation at the detector. In Fig. 5-6A, two triodes are used as cathode followers, for the signal and for the b.f.o., working into a common cathode resistor (1000 ohms). The third triode also shares this cathode resistor and consequently the same signals, but it has an audio load in its plate circuit and it operates at a higher grid bias (by virtue of the 2700-ohm resistor in its cathode circuit). The signals and the b.f.o. mix in this third triode. If the b.f.o. is turned off, a modulated signal running through the signal cathode follower should yield little or no audio output from the detector, up to the overload point of the signal cathode follower. Turning on the b.f.o. brings in modulation, because now the detector output is the product of the two signals. The plates of the cathode followers are grounded and filtered for the i.f., and the 4700-μμf. capacitor from plate to ground in the output triode furnishes a bypass at the i.f. The b.f.o. voltage should be about 2 r.m.s., and the signal should not exceed about 0.3 volts r.m.s.

The circuit in Fig. 5-6B is a simplification requiring one less triode. Its principle of operation is substantially the same except that the additional bias for the output tube is derived from rectified b.f.o. voltage across the 100,000-ohm resistor. More elaborate r.f. filtering is shown in the plate of the output tube (2-mh. choke and the 220-μμf. capacitors), and the degree of plate filtering in either circuit will depend upon the frequencies involved. At low intermediate frequencies, more elaborate filtering is required.

● REGENERATIVE DETECTORS

By providing controllable r.f. feedback (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 thus the selectivity. The grid-leak type of detector is most suitable for the purpose.

The grid-leak detector is a combination diode rectifier and audio-frequency amplifier. In the circuit of Fig. 5-7A, 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, and the audio-frequency variations in voltage across R_1 are amplified through the tube as in a normal a.f. amplifier. In the plate circuit, R_2 is the plate load resistance and C_3 and RFC a filter to eliminate r.f. in the output circuit.

A grid-leak detector has considerably greater sensitivity than a diode. The sensitivity is further increased by using a screen-grid tube instead of a triode. The operation is equivalent to that of the triode circuit. The screen bypass capacitor should have low reactance for both radio and audio frequencies.

The circuit in Fig. 5-7B is regenerative, the feedback being obtained by feeding some signal from the plate circuit back to the grid by inductive coupling. 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. The critical

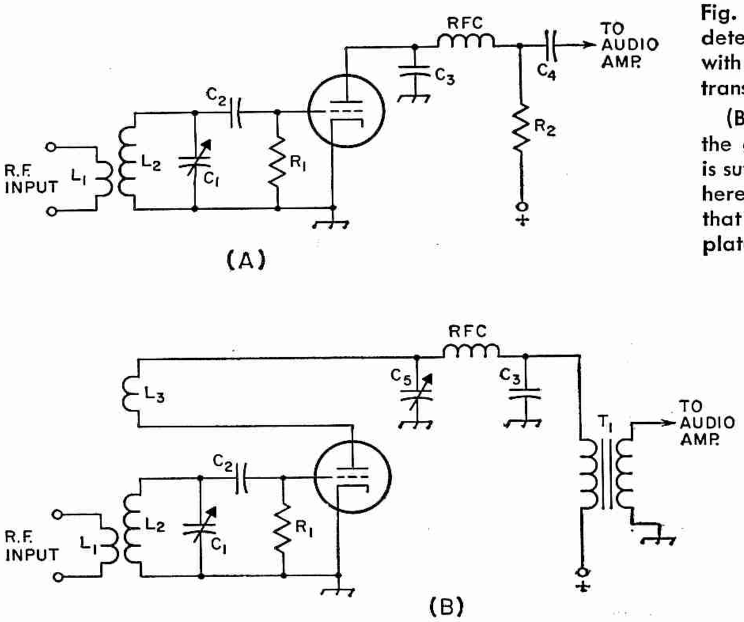


Fig. 5-7—(A) Triode grid-leak detector combines diode detection with triode amplification. Although shown here with resistive plate load, R_2, an audio choke coil or transformer could be used.

(B) Feeding some signal from the plate circuit back to the grid makes the circuit regenerative. When feedback is sufficient, the circuit will oscillate. Feedback is controlled here by varying reactance at C_5; with fixed capacitor at that point regeneration could be controlled by varying plate voltage or coupling between L_2 and L_3.

point in turn depends upon circuit conditions, which may vary with the frequency to which the detector is tuned. An oscillating detector can be detuned slightly from an incoming c.w. signal to give *autodyne* reception.

The circuit of Fig. 5-7B uses a variable bypass capacitor, C_5, in the plate circuit to control regeneration. When the capacitance is small the tube does not regenerate, but as it increases toward maximum its reactance becomes smaller until there is sufficient feedback 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 end of L_2.

Although the regenerative grid-leak detector is more sensitive than any other type, its many disadvantages 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.

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 will result in a slight decrease in the hiss.

The proper adjustment of the regeneration control for best reception of code signals is where the detector just starts to oscillate. Then code 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" and rise again on the other side, finally disappearing at a very high pitch. This behavior is shown in Fig. 5-8. A low-pitched beat-note cannot be obtained from a strong signal because the detector "**pulls in**" or "**blocks**"; that is, the signal forces the detector to oscillate at the signal frequency, even though the circuit may not be tuned exactly to the signal. It usually can be corrected by advancing the regeneration control until the beat-note is heard again, or by reducing the input signal.

The point just after the detector starts oscillating is the most sensitive condition for code reception. Further advancing the regeneration control makes the receiver less prone to blocking, 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.

Single-sideband phone signals can be received with a regenerative detector by advancing the regeneration control to the point used for code reception and tuning carefully across the s.s.b. signal. The tuning will be very critical, however, and the operator must be prepared to just "creep" across the signal. A strong signal will pull the detector and make reception impossible, so either the regeneration must be advanced far enough to prevent this condition, or the signal must be reduced by using loose antenna coupling.

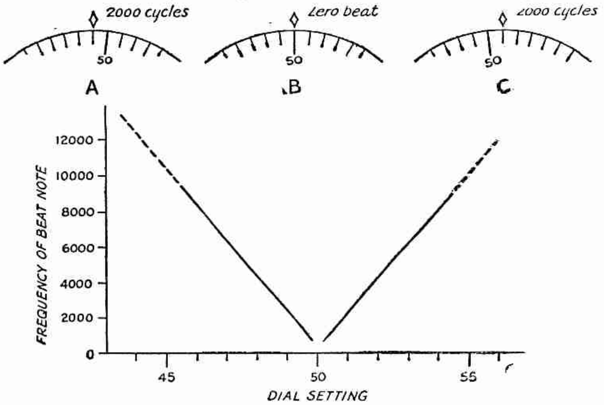


Fig. 5-8—As the tuning dial of a receiver is turned past a code signal, the beat-note varies from a high tone down through "zero beat" (no audible frequency difference) and back up to a high tone, as shown at A, B and C. The curve is a graphical representation of the action. The beat exists past 8000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

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 capacitor 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 capacitor 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. This is not necessary if the coils are separated from each other by several coil diameters, or are mounted at right angles to each other.

Another method is to use coils wound on forms with contacts (usually pins) that can be plugged in and removed from a socket. These plug-in 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 capacitor will depend upon the inductance of the coil and the change in tuning capacity. For ease of tuning, it is desirable to adjust the tuning range so that practically the whole dial scale is occupied by the band in use. This is called **bandspreading.** Because of the varying widths of the bands, special tuning methods must be devised to give the correct maximum-minimum capacity ratio on each band. Several of these methods are shown in Fig. 5-9.

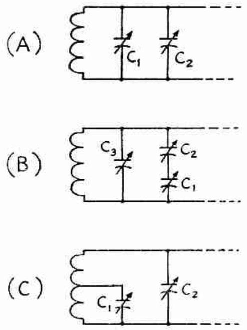


Fig. 5-9—Essentials of the three basic bandspread tuning systems.

In A, a small **bandspread capacitor,** C_1 (15- to 25-$\mu\mu$f. maximum capacity), is used in parallel with a capacitor, C_2, which is usually large enough (100 to 140 $\mu\mu$f.) to cover a 2-to-1 frequency range. The setting of C_2 will determine the minimum capacitance 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. It is almost impossible, because of the nonharmonic relation of the various band limits, to get full bandspread on all bands with the same pair of capacitors. C_2 is variously called the **band-setting** or **main-tuning** capacitor. It must be reset each time the band is changed.

The method shown at B makes use of capacitors in series. The tuning capacitor, C_1, may have a maximum capacitance of 100 $\mu\mu$f. or more. The minimum capacitance is determined principally by the setting of C_3, which usually has low capacitance, and the maximum capacitance by the setting of C_2, which is of the order of 25 to 50 $\mu\mu$f. 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 capacitors must be switched in.

The circuit at C also gives complete spread on each band. C_1, the bandspread capacitor, may have any convenient value; 50 $\mu\mu$f. is satisfactory. C_2 may be used for continuous frequency coverage ("general coverage") and as a band-setting capacitor. The effective maximum-minimum capacitance ratio depends upon 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 capacitance. C_2 may be connected permanently across the individual inductor and preset, if desired. This requires a separate capacitor for each band, but eliminates the necessity for resetting C_2 each time.

Ganged Tuning

The tuning capacitors 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 capacitors, and circuit inductances and minimum and maximum capacities are identical in all "ganged" stages. A small **trimmer** or **padding** capacitor may be connected across the coil, so that variations in minimum capacity can be compensated. The fundamental circuit is shown in Fig. 5-10, where C_1 is the trimmer and C_2 the tuning capacitor. The use of the trimmer necessarily increases the minimum circuit capacity, but it is a necessity for satisfactory tracking. Midget capacitors having maximum capacities of 15 to 30 μμf. are commonly used.

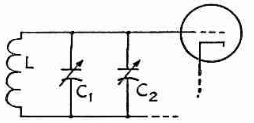


Fig. 5-10—Showing the use of a trimmer capacitor to set the minimum circuit capacity in order to obtain true tracking for gang-tuning.

The same methods are applied to bandspread circuits that must be tracked. The circuits are identical with those of Fig. 5-9. If both general-coverage and bandspread tuning are to be available, an additional trimmer capacitor must be connected across the coil in each circuit shown. If only amateur-band tuning is desired, however, then C_3 in Fig. 5-9B, and C_2 in Fig. 5-9C, serve as trimmers.

The coil inductance can be adjusted by starting with a larger number of turns than necessary and removing a turn or fraction of a turn at a time until the circuits track satisfactorily. An alternative method, provided the inductance is reasonably close to the correct value initially, is to make the coil so that the last turn is variable with respect to the whole coil.

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 raise the Q of a coil, provided the iron is 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 (until 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 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 heterodyned to a new radio frequency, the **intermediate frequency** (abbreviated **"i.f."**), then amplified, and finally detected. The frequency is changed by modulating the output of a tunable oscillator (the **high-frequency,** or **local, oscillator**) by the incoming signal in a **mixer** or **converter** stage (**first detector**) to produce a side frequency equal to the intermediate frequency. The other side frequency is rejected by selective circuits. The audio-frequency signal is obtained at the **second detector.** Code 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 at 7000 kc. Then the high-frequency oscillator frequency may be set to 7455 kc., in order that one side 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 code 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 good stability and, since they are working at frequencies considerably removed from the signal frequencies (percentage-wise), they are not normally "pulled" 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 signal is heterodyned 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 a low power 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

A circuit tuned to the intermediate frequency is placed in the plate circuit of the mixer, to offer a high impedance load for the i.f. voltage that is developed. The signal- and oscillator-frequency voltages appearing in the plate circuit are rejected by the selectivity of this circuit. 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.

A change in oscillator frequency caused by tuning of the mixer grid circuit is called **pulling.** 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 voltage to change, which 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.

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

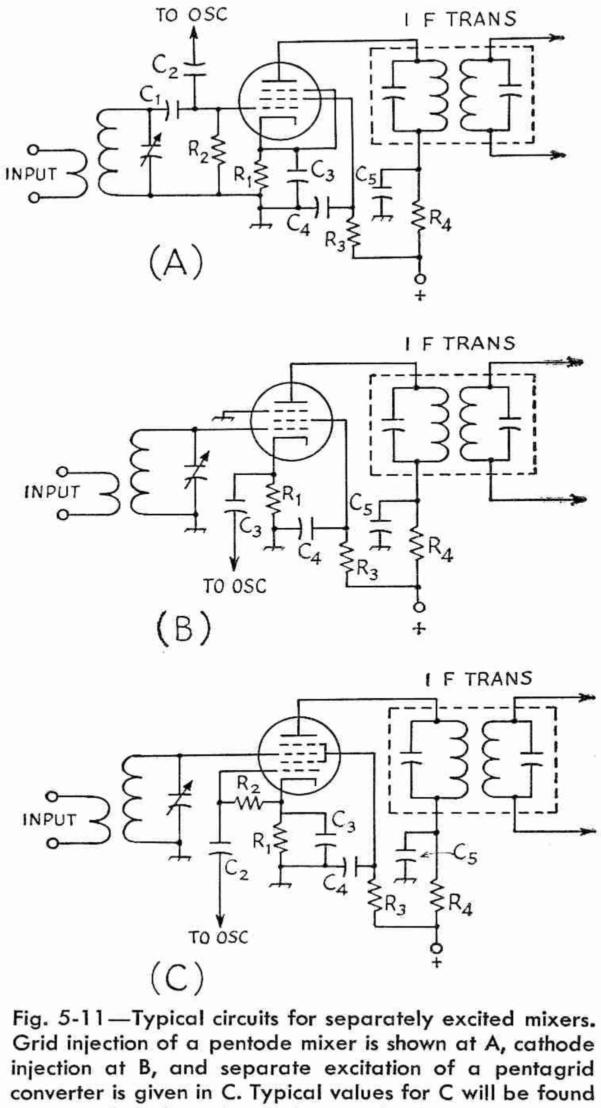


Fig. 5-11—Typical circuits for separately excited mixers. Grid injection of a pentode mixer is shown at A, cathode injection at B, and separate excitation of a pentagrid converter is given in C. Typical values for C will be found in Table 5-I—the values below are for the pentode mixer of A and B.

C_1—10 to 50 μμf.
C_2—5 to 10 μμf.
C_3, C_4, C_5—0.001 μf.
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 or 6AH6, 150 with a 6AK5.

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-transconductance tubes like the 6AC7, 6AK5 or 6U8 (pentode section). Triode tubes can be used as mixers in grid-injection circuits, but they are commonly used only at 50 Mc. and higher, where mixer noise may become a significant factor. The triode mixer has the lowest inherent noise, the pentode is next, and the multigrid converter tubes are the noisiest.

The circuit in Fig. 5-11B shows cathode injection at the mixer. Operation is similar to the grid-injection case, and the same considerations apply.

It is difficult to avoid "pulling" in a triode or pentode mixer, and a pentagrid mixer tube provides much better isolation. A typical circuit is shown in Fig. 5-11C, and tubes like the 6SA7, 6BA7 or 6BE6 are commonly used. The oscillator voltage is introduced 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 much noisier than a triode or pentode mixer, but its isolating characteristics make it a very useful device.

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, 6SB7Y, 6BA7 or 6BE6. Generally the only care necessary is to adjust the feedback 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.

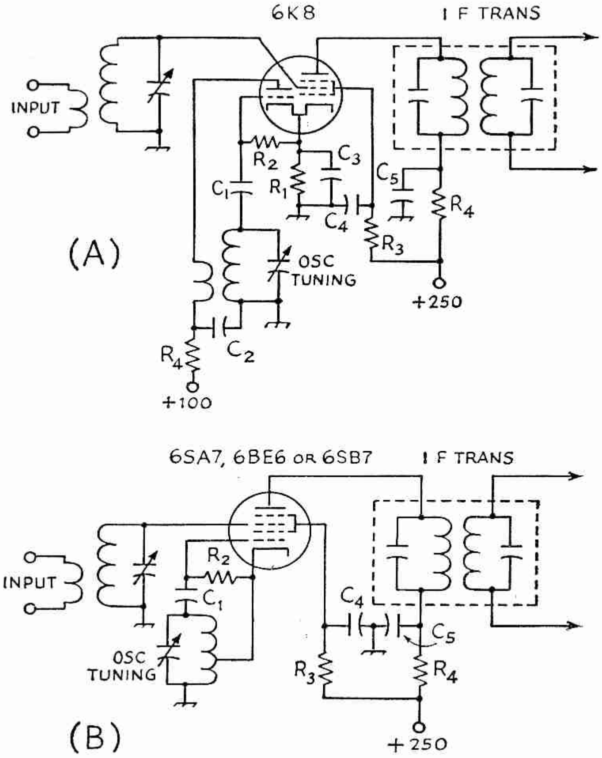


Fig. 5-12—Typical circuits for triode-hexode (A) and pentagrid (B) converters. Values for R_1, R_2 and R_3 can be found in Table 5-I; others are given below.

C_1—47 μμf. **C_3—0.01 μf.**
C_2, C_4, C_5—0.001 μf. **R_4—1000 ohms.**

A more stable receiver generally results, particularly at the higher frequencies, when separate tubes are used for the mixer and oscillator. Practically the same number of circuit components is required whether or not a combination tube is used, so that there is very little difference to be realized from the cost standpoint.

Typical circuit constants for converter tubes are given in Table 5-I. The grid leak referred to is the oscillator grid leak or injection-grid return, R_2 of Figs. 5-11C 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

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]	240	27,000	47,000	0.15-0.2	—	—	—	—
6SA7[2]	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.

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. 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 cut down the gain. For this reason, and to reduce image response, the i.f. following the first converter of a receiver should be not less than 5 or 10 per cent of the signal frequency, for best results.

Transistors in Mixers

Typical transistor circuitry for a mixer operating at frequencies below 20 Mc. is shown in Fig. 5-13. The local oscillator current is injected in the emitter circuit by inductive coupling to L_1; L_1 should have low reactance at the oscillator frequency. The input from the r.f. amplifier should be at low impedance, obtained by inductive coupling or tapping down on the tuned circuit. The output transformer T_1 has the collector connection tapped down on the inductance to maintain a high Q in the tuned circuit.

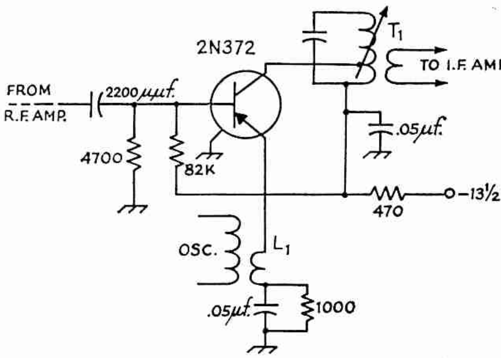


Fig. 5-13--Typical transistor mixer circuit.

L_1—Low-impedance inductive coupling to oscillator.

T_1—Transistor i.f. transformer. Primary impedance of 100,000 ohms, secondary impedance of 1700 ohms, unloaded $Q = 100$, loaded $Q = 35$.

Audio Converters

Converter circuits of the type shown in Fig. 5-12 can be used to advantage in the reception of code 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 "product detector" of Fig. 5-6 is also a converter circuit, and the statements above for audio converters apply to the product detector.

THE HIGH-FREQUENCY OSCILLATOR

Stability of the receiver is dependent chiefly upon the stability of the h.f. oscillator, and particular care should be given this part of the receiver. The frequency of oscillation should be insensitive to mechanical shock and changes in voltage and loading. Thermal effects (slow change in frequency because of tube or circuit heating) should be minimized. They can be reduced by using ceramic instead of bakelite insulation in the r.f. circuits, a large cabinet relative to the chassis (to provide for good radiation of developed heat), minimizing the number of high-wattage resistors in the receiver and putting them in the separate power supply, and not mounting the oscillator coils and tuning capacitor 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 minimized by using good mechanical support for coils and tuning capacitors, a heavy chassis, and by not hanging any of the oscillator-circuit components on long leads. Tie-points should be used to avoid long leads. 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 capacitors. They should have good alignment and no back-lash. If the capacitors 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 capacitors 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, because 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.

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 responses.

The oscillator plate power should be as low as is consistent with adequate output. Low plate power will reduce tube heating and thereby lower the frequency drift. The oscillator and mixer circuits should be well isolated, pref-

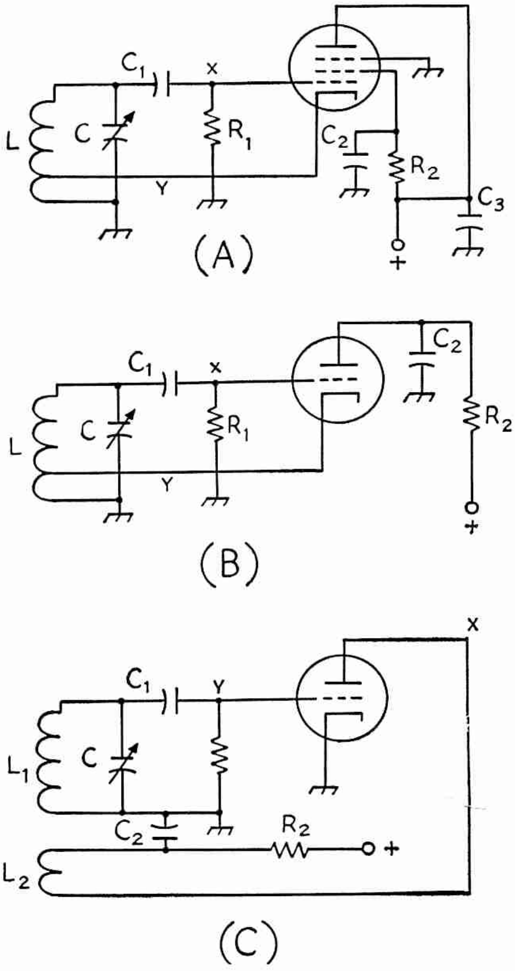


Fig. 5-14—High-frequency oscillator circuits. A, pentode grounded-plate oscillator; B, triode grounded-plate oscillator; C, triode oscillator with tickler circuit. Coupling to the mixer may be taken from points X and Y. In A and B, coupling from Y will reduce pulling effects, but gives less voltage than from X; this type is best adapted to mixer circuits with small oscillator-voltage requirements. Typical values for components are as follows:

	Circuit A	Circuit B	Circuit C
C_1	100 μμf.	100 μμf.	100 μμf.
C_2	0.01 μf.	0.01 μf.	0.01 μf.
C_3	0.01 μf.		
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.

erably by shielding, since coupling other than by the intended means may result in pulling.

If the h.f.-oscillator frequency is affected by changes in plate voltage, a voltage-regulated plate supply (VR tube) can be used.

Circuits

Several oscillator circuits are shown in Fig. 5-14. 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. 5-14C reduces hum because the cathode is grounded. It is simple to adjust, and it is also the best circuit to use with filament-type tubes. With filament-type tubes, the other two circuits would require r.f. chokes to keep the filament above r.f. ground.

Besides the use of a fairly high C/L ratio in the tuned circuit, it is necessary to adjust the feedback to obtain optimum results. Too much feedback may cause **"squegging"** of the oscillator and the generation of several frequencies simultaneously; too little feedback will cause the output to be low. In the tapped-coil circuits (A, B), the feedback is increased by moving the tap toward the grid end of the coil. In C, more feedback is obtained by increasing the number of turns on L_2 or 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 elaborate sets.

Choice of Frequency

The selection of an intermediate frequency is a compromise between 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 gain is lowered and selectivity is harder to obtain by simple means.

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 without very loose coupling between mixer and oscillator.

With an i.f. of about 1600 kc., satisfactory image ratios can be secured on 14, 21 and 28 Mc. with one r.f. stage of good design. For frequencies of 28 Mc. and higher, a common solution is to use a double superheterodyne, choosing one high i.f. for image reduction (5 and 10 Mc. are frequently used) and a lower one for gain and selectivity.

In choosing an i.f. it is wise to avoid frequencies on which there is considerable activity by the various radio services, since such signals may be picked up directly on the i.f. wiring. Shifting the i.f. or better shielding are the solutions to this 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, the i.f. 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 is 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.

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 in crowded amateur 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. 5-15. A second stage would simply duplicate the circuit of the first. The i.f. amplifier practically always uses a remote cut-off pentode-type tube operated as a Class A amplifier. For maximum selectivity, double-tuned transformers are used for interstage coupling, although single-tuned circuits or transformers with untuned primaries can be used for coupling, with a consequent loss in selectivity. All other things being equal, the selectivity of an i.f. amplifier is proportional to the number of tuned circuits in it.

In Fig. 5-15, the gain of the stage is reduced by introducing a negative voltage to the lead marked "AGC" or a positive voltage to R_1 at the point marked "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 prevent unwanted interstage coupling. C_2 and R_4 are part of the automatic gain-control circuit (described later); if no a.g.c. is used, the lower end of the i.f.-transformer secondary is connected to chassis.

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 normally has no effect on the signal-to-noise ratio, since this is determined by the preceding mixer and r.f. amplifier.

Typical values of cathode and screen resistors for common tubes are given in Table 5-II. The 6K7, 6SK7 and 6BJ6 are recommended for i.f. work because they have desirable remote cut-off characteristics. The indicated screen resistors drop the plate voltage to the correct screen voltage, as R_2 in Fig. 5-15.

When two or more stages are used the high gain may tend to cause instability and oscillation, so that good shielding, bypassing, and careful circuit arrangement to prevent stray coupling between input and output circuits are necessary.

When single-ended tubes are used, the plate and grid leads should be well separated. With these tubes it is advisable to mount the screen

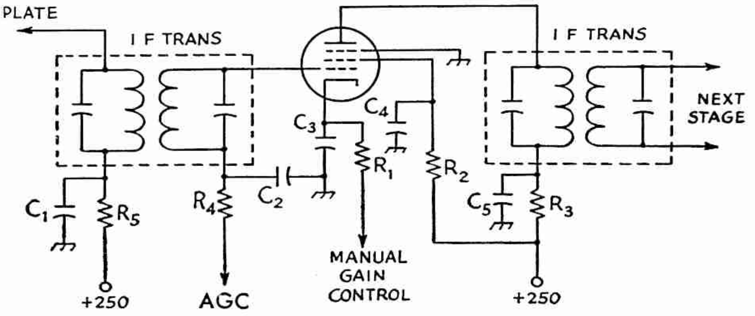


Fig. 5-15—Typical intermediate-frequency amplifier circuit for a superheterodyne receiver. Representative values for components are as follows:

C_1, C_3, C_4, C_5—0.02 μf. at 455 kc.; 0.01 μf. at 1600 kc. and higher.

C_2—0.01 μf.

R_1, R_2—See Table 5-II.

R_3, R_5—1500 ohms.

R_4—0.1 megohm.

TABLE 5-II

Cathode and Screen-Dropping Resistors for R.F. or I.F. Amplifiers

Tube	*Plate Volts*	*Screen Volts*	*Cathode Resistor* R1	*Screen Resistor* R2
6AC7¹	300		160	62,000
6AH6²	300	150	160	62,000
6AK5²	180	120	200	27,000
6AU6²	250	150	68	33,000
6BA6²*	250	100	68	33,000
6BH6²	250	150	100	33,000
6BJ6²*	250	100	82	47,000
6BZ6²*	200	150	180	20,000
6CB6	200	150	180	56,000
6SG7¹*	250	125	68	27,000
6SH7¹	250	150	68	39,000
6SJ7¹	250	100	820	180,000
6SK7¹*	250	100	270	56,000

¹ Octal base, metal. ² Miniature tube

* Remote cut-off type.

bypass capacitor directly on the bottom of the socket, crosswise between the plate and grid pins, to provide additional shielding. If a paper capacitor is used, the outside foil should be grounded to the chassis.

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 capacitors 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. 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 capacitance can exist between layers. Universal winding, with its "criss-crossed" turns, tends to reduce distributed-capacity effects.

For tuning, air-dielectric tuning capacitors are preferable to mica compression types 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-capacitor tuning can be obtained by use of high-stability fixed mica or ceramic capacitors. Such stability is of great importance, since a circuit whose frequency "drifts" with time eventually will be tuned to a different frequency than the other circuits, thereby reducing the gain and selectivity of the amplifier. Typical i.f.-transformer construction is shown in Fig. 5-16.

The normal **interstage** i.f. transformer is loosely coupled, to give good selectivity consistent with adequate gain. A so-called **diode transformer** is similar, but the coupling is tighter, to give sufficient transfer when working into the finite load presented by a diode detector. Using a diode transformer in place of an interstage transformer would result in loss of selectivity; using an interstage transformer to couple to the diode would result in loss of gain.

Besides the type of i.f. transformer shown in Fig. 5-16, special units to give desired selectivity characteristics are available. For higher-than-ordinary adjacent-channel selectivity **triple-tuned** transformers, with a third tuned circuit inserted between the input and output windings, are sometimes 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.

A method of varying the selectivity is to vary the coupling between primary and secondary, overcoupling being used to broaden the selectivity curve. Special circuits using single tuned circuits, coupled in any of several different ways, are used in some advanced receivers.

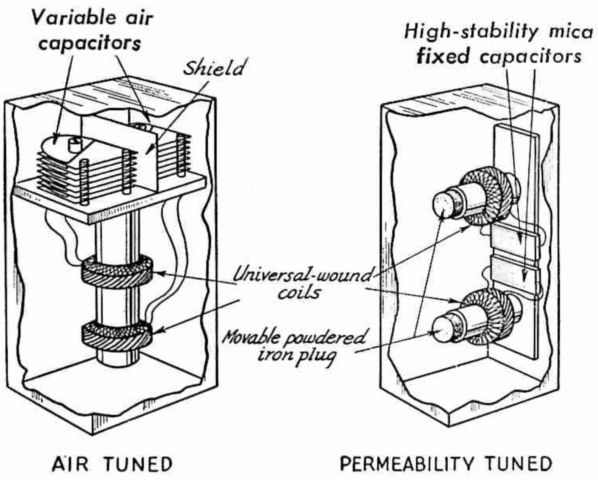


Fig. 5-16—Representative i.f.-transformer construction. Coils are supported on insulating tubing or (in the air-tuned type) on wax-impregnated wooden dowels. The shield in the air-tuned transformer prevents capacity coupling between the tuning capacitors. In the permeability-tuned transformer the cores consist of finely-divided iron particles supported in an insulating binder, formed into cylindrical "plugs." The tuning capacitance is fixed, and the inductances of the coils are varied by moving the iron plugs in and out.

Selectivity

The over-all selectivity of the r.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	*6 db. down*	*20 db. down*	*40 db. down*
One stage, 50 kc. (iron core)	2.0	3.0	4.2
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

Transistor I. F. Amplifier

A typical circuit for a two-stage transistor i.f. amplifier is shown in Fig. 5-17. Constants are given for a 455-kc. amplifier, but the same gen-

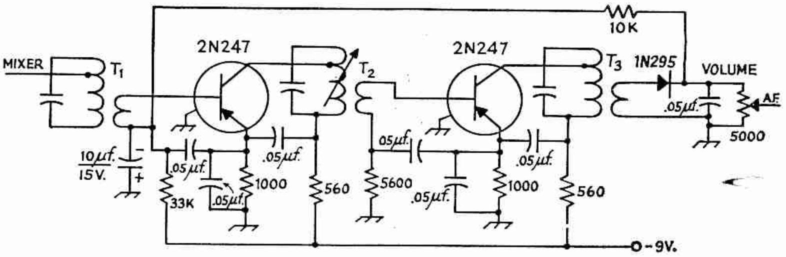


Fig. 5-17—Typical circuit for a two-stage transistor i.f. amplifier. At high frequencies a neutralizing capacitor may be required, as mentioned in the text.

T_1—Transistor input i.f. transformer. Primary impedance = 100,000 ohms, secondary impedance = 1700 ohms, unloaded Q = 100, loaded Q = 35.

T_2—Transistor interstage i.f. transformer. Primary impedance = 4600 ohms, secondary impedance = 1700 ohms, unloaded Q = 39, loaded Q = 35.

T_3—Transistor output i.f. transformer. Primary impedance = 30,000 ohms, secondary impedance = 1000 ohms, unloaded Q = 100, loaded Q = 35.

eral circuitry applies to an amplifier at any frequency within the operating range of the transistors. When higher frequencies are used, it may be necessary to neutralize the amplifier to avoid overall oscillation; this is done by connecting a small variable capacitor of a few $\mu\mu$f. from base to base of the transistors.

Automatic gain control is obtained by using the developed d.c. at the 1N295 diode detector to modify the emitter bias current on the first stage. As the bias current changes, the input and output impedances change, and the resultant impedance mismatches causes a reduction in gain. Such a.g.c. assumes, of course, that the amplifier is set up initially in a matched condition.

● THE SECOND DETECTOR AND BEAT OSCILLATOR

Detector Circuits

The second detector of a superheterodyne receiver 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

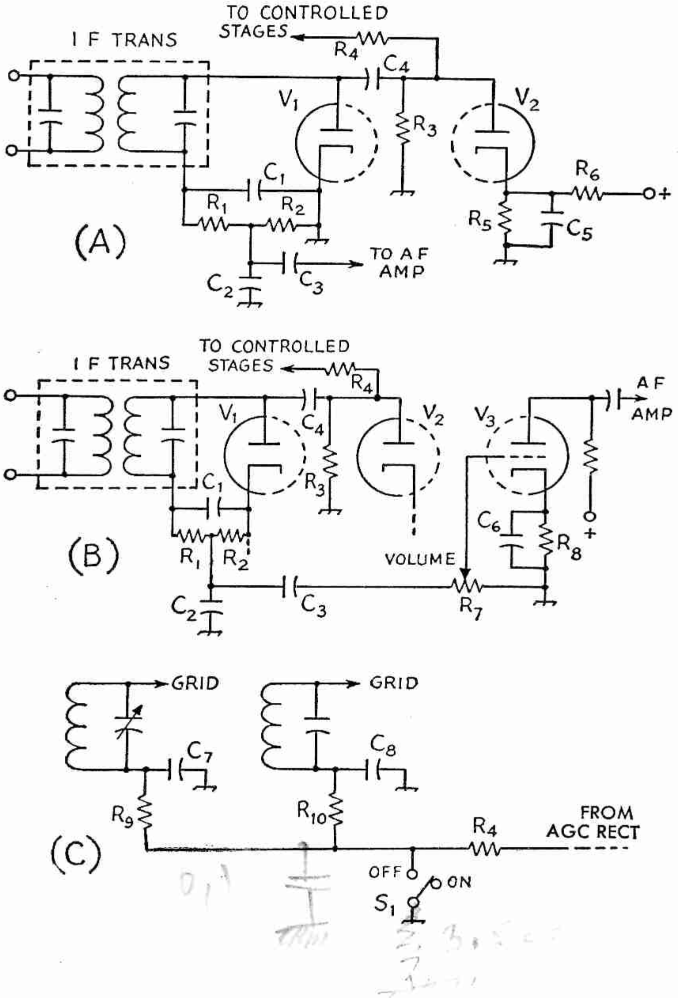


Fig. 5-18—Delayed automatic gain-control circuits using a twin diode (A) and a dual-diode triode. The circuits are essentially the same and differ only in the method of biasing the a.g.c. rectifier. The a.g.c. control voltage is applied to the controlled stages as in (C). For these circuits typical values are:

C_1, C_3, C_4—100 $\mu\mu$f.
C_2, C_5, C_7, C_8—0.01 μf.
C_6—5-μf. electrolytic.
R_1, R_9, R_{10}—0.1 megohm.
R_2—0.47 megohm.
R_3—2 megohms.
R_4—0.47 megohm.
R_5, R_6—Voltage divider to give 2 to 10 volts bias at 1 to 2 ma. drain.
R_7—0.5-megohm volume control.
R_8—Correct bias resistor for triode section of dual-diode triode.

cases the diode elements are incorporated in a multipurpose tube that contains an amplifier section in addition to the diode.

Audio-converter circuits and product detectors are often used for code or s.s.b. detectors.

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 the circuits shown in Fig. 5-14A and B, with the output taken from Y. A variable capacitor of about 25-$\mu\mu$f. capacitance can be connected between cathode and ground to provide fine adjustment of the frequency. The beat oscillator usually is coupled to the second-detector tuned circuit through a fixed capacitor of a few $\mu\mu$f.

The beat oscillator should be well shielded, to prevent coupling to any part of the receiver except the second detector and to prevent its harmonics from getting into the front end and being amplified along 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.

AUTOMATIC GAIN CONTROL

Automatic regulation of the gain of the receiver in inverse proportion to the signal strength is an operating convenience 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 and the output more constant as the number of stages to which the a.g.c. bias is applied is increased. Control of at least two stages is advisable.

Circuits

Although some receivers derive the a.g.c. voltage from the diode detector, the usual practice is to use a separate a.g.c. rectifier. Typical circuits are shown in Figs. 5-18A and 5-18B. The two rectifiers can be combined in one tube, as in the 6H6 and 6AL5. In Fig. 5-18A V_1 is the diode detector; the signal is developed across R_1R_2 and coupled to the audio stages through C_3. C_1, R_1 and C_2 are included for r.f. filtering, to prevent a large r.f. component being coupled to the audio circuits. The a.g.c. rectifier, V_2, is coupled to the last i.f. transformer through C_4, and most of the rectified voltage is developed across R_3. V_2 does not rectify on weak signals, however; the fixed bias at R_5 must be exceeded before rectification can take place. The developed negative a.g.c. bias is fed to the controlled stages through R_4.

The circuit of Fig. 5-18B is similar, except that a dual-diode triode tube is used. Since this has only one common cathode, the circuitry is slightly different but the principle is the same. The triode stage serves as the first audio stage, and its bias is developed in the cathode circuit across R_8. This same bias is applied to the a.g.c. rectifier by returning its load resistor, R_3, to ground. To avoid placing this bias on the detector, V_1, its load resistor R_1R_2 is returned to cathode, thus avoiding any bias on the detector and permitting it to respond to weak signals.

The developed negative a.g.c. bias is applied to the controlled stages through their grid circuits, as shown in Fig. 5-18C. C_7R_9 and C_8R_{10} serve as filters to avoid common coupling and possible feedback and oscillator. The a.g.c. is disabled by closing switch S_1.

The a.g.c. rectifier bias in Fig. 5-18B is set by the bias required for proper operation of V_3. If less bias for the a.g.c. rectifier is required, R_3 can be tapped up on R_8 instead of being returned to chassis ground. In Fig. 5-18A, proper choice of bias at R_5 depends upon the over-all gain of the receiver and the number of controlled stages. In general, the bias at R_5 will be made higher for receivers with more gain and more stages.

Time Constant

The time constant of the resistor-capacitor combinations in the a.g.c. circuit is an important part of the system. It must be long 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.g.c. voltage applied to the amplifier grids would reduce the percentage of modulation on the incoming signal. But the time constant must not be too long or the a.g.c. will be unable to follow rapid fading. The capacitance and resistance values indicated in Fig. 5-18 will give a time constant that is satisfactory for average reception.

C.W. and S.S.B.

A.g.c. can be used for c.w. and s.s.b. reception but the circuit is usually more complicated. The a.g.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 done by using a separate a.g.c. channel connected to an i.f. amplifier stage ahead of the second detector (and b.f.o.) or by rectifying the audio output of the detector. If the selectivity ahead of the a.v.c. rectifier isn't good, strong adjacent-channel signals may develop a.g.c. voltages that will reduce the receiver gain while listening to weak signals. When clear channels are available, however, c.w. and s.s.b. a.g.c. will hold the receiver output constant over

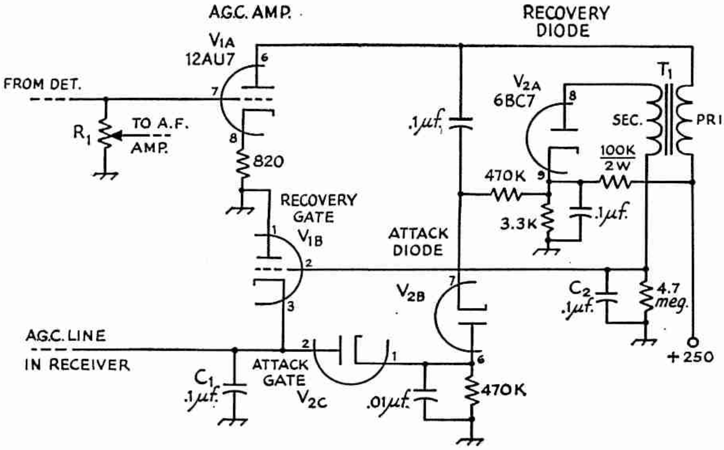


Fig. 5-19—Audio "hang" a.g.c. system. Resistors are ½-watt unless specified otherwise.

R1—Normal audio volume control in receiver.

T1—1:3 step-up audio transformer (Stancor A-53 or equiv.)

The hang time can be adjusted by changing the value of the recovery diode load resistor (4.7 megohms shown here). The a.g.c. line in the receiver must have no d.c. return to ground and the receiver should have good skirt selectivity for maximum effectiveness at the system.

a wide range of signal inputs. A.g.c. systems designed to work on these signals should have fast-attack and slow-decay characteristics to work satisfactorily, and often a selection of time constants is made available.

The a.g.c. circuit shown in Fig. 5-19 is applicable to many receivers without too much modification. Audio from the receiver is amplified in V_{1A} and rectified in V_{2B}. The resultant voltage is applied to the a.g.c. line through V_{2C}. The capacitor C_1 charges quickly and will remain charged until discharged by V_{1B}. This will occur some time after the signal has disappeared, because the audio was stepped up through T_1 and rectified in V_{2A}, and the resultant used to charge C_2. This voltage holds V_{1B} cut off for an appreciable time, until C_2 discharges through the 4.7-megohm resistor. The threshold of compression is set by adjusting the bias on the diodes (changing the value of the 3.3K or 100K resistors). There can be no d.c. return to ground from the a.g.c. line, because C_1 must be discharged only by V_{1B}. Even a v.t.v.m. across the a.g.c. line will be too low a resistance, and the operation of the system must be observed by the action of the S meter.

Occasionally a strong noise pulse may cause the a.g.c. to hang until C_2 discharges, but most of the time the gain should return very rapidly to that set by the signal. A.g.c. of this type is very helpful in handling netted s.s.b. signals of widely varying 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 or industrial 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 short duration of the pulses compared with the time between them, must have high 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 principles of devices intended to reduce such noise is to allow the desired signal to pass through the receiver unaffected, but to make 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.

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 of the circuits. Thus the more selectivity ahead of the noise-reducing device, the more difficult it becomes to secure good pulse-type 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

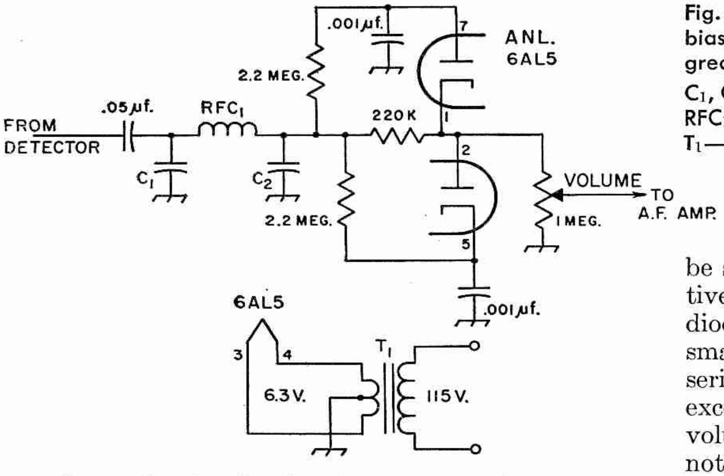


Fig. 5-20—Full-wave shunt limiter using contact-potential-biased diodes. A low-level limiter (½ volt), this circuit finds greatest usefulness following a product detector.

C_1, C_2—Part of low-pass filter with cutoff below i.f.
RFC_1—Part of low-pass filter; see C_1.
T_1—Center-tapped heater transformer.

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 stages.

SECOND-DETECTOR NOISE LIMITER CIRCUITS

Most audio limiting circuits are based on one of two principles. In a series limiting circuit, a normally conducting element (or elements) is connected in the circuit in series and operated in such a manner that it becomes non-conductive above a given signal level. In a shunt limiting circuit, a non-conducting element is connected in shunt across the circuit and operated so that it becomes conductive above a given signal level, thus short-circuiting the signal and preventing its being transmitted to the remainder of the amplifier. The usual conducting element will be a forward-biased diode, and the usual non-conducting element will be a back-biased diode. In many applications the value of bias is set manually by the operator; usually the clipping level will be set at about 5 to 10 volts.

A full-wave clipping circuit that operates at a low level (approximately ½ volt) is shown in Fig. 5-20. Each diode is biased by its own contact potential, developed across the 2.2-megohm resistors. The .001-μf. capacitors become charged to close to this value of contact potential. A negative-going signal in excess of the bias will be shorted to ground by the upper diode; a positive-going signal will be conducted by the lower diode. The conducting resistance of the diodes is small by comparison with the 220,000 ohms in series with the circuit, and little if any of the excessive signal will appear across the 1-megohm volume control. In order that the clipping does not become excessive and cause distortion, the input signal must be held down by a gain control ahead of the detector. This circuit finds good application following a low-level detector.

To minimize hum in the receiver output, it is desirable to ground the center tap of the heater transformer, as shown, instead of the more common practice of returning one side of the heater circuit to chassis.

Second-detector noise-limiting circuits that automatically adjust themselves to the received carrier level are shown in Fig. 5-21. 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. bypass. 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 long time constant of C_2R_3 prevents any rapid change of the 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. Diode rectifiers such as the 6H6 and 6AL5 can be used for these types of noise limiters. Neither circuit is useful for c.w. or s.s.b. reception, but they are both quite effective

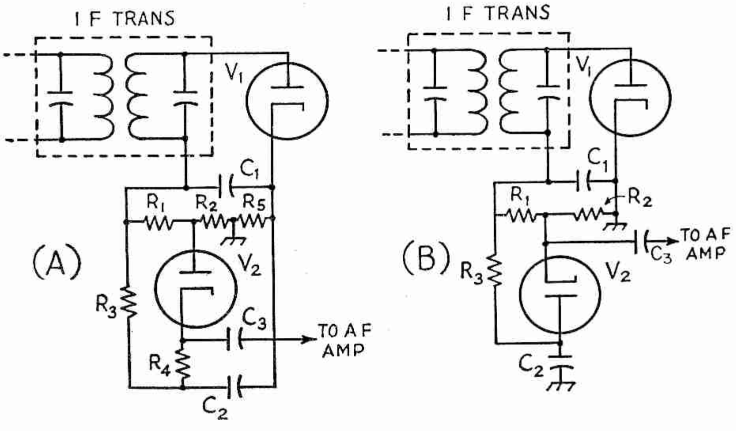


Fig. 5-21—Self-adjusting series (A) and shunt (B) noise limiters. The functions of V_1 and V_2 can be combined in one tube like the 6H6 or 6AL5.

C_1—100 μμf.
C_2, C_3—0.05 μf.
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.
R_5—6800 ohms.

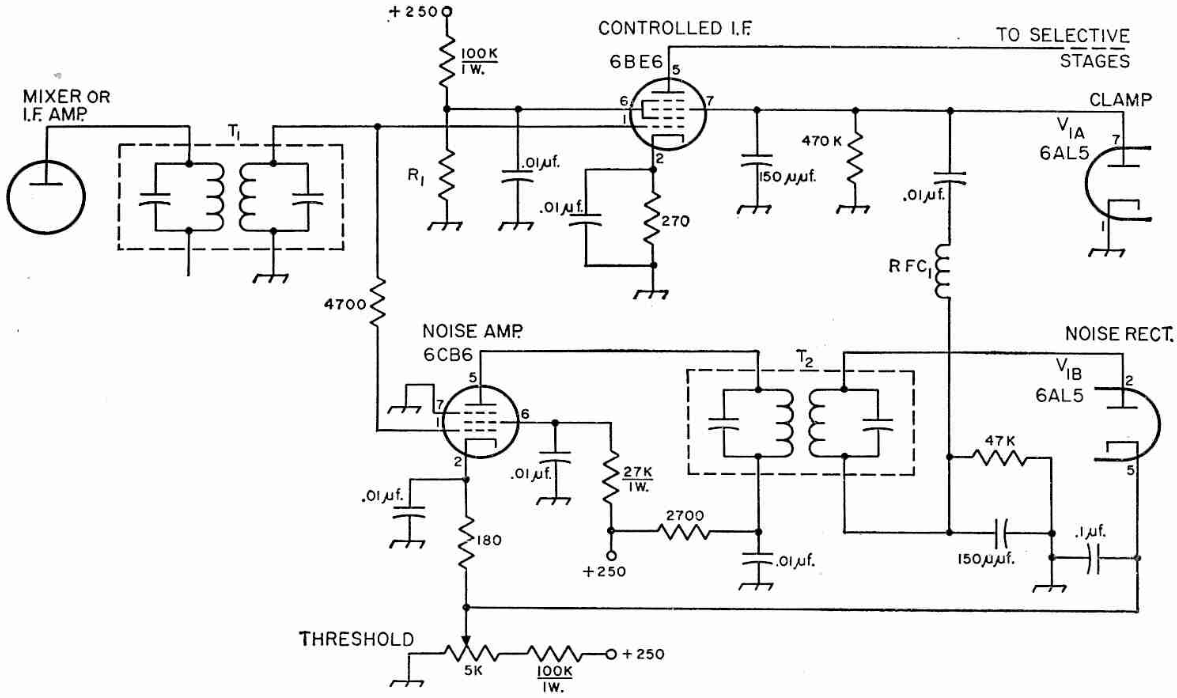


Fig. 5-22—Practical circuit diagram of an i.f. noise silencer. For best results the silencer should be used ahead of the high-selectivity portion of the receiver.

T_1—Interstage i.f. transformer

T_2—Diode i.f. transformer.

R_1—33,000 to 68,000 ohms, depending upon gain up to this stage.

RFC_1—R.f. choke, preferably self-resonant at i.f.

for a.m. phone work. The series circuit (A) is slightly better than the shunt circuit.

I.F. NOISE SILENCER

The i.f. noise silencer circuit shown in Fig. 5-22 is designed to be used in a receiver as far along from the antenna stage as possible but ahead of the high-selectivity section of the receiver. Noise pulses are amplified and rectified, and the resulting negative-going d.c. pulses are used to cut off an amplifier stage during the pulse. A manual "threshold" control is set by the operator to a level that only permits rectification of the noise pulses that rise above the peak amplitude of the desired signal. The clamp diode, V_{1A}, short circuits the positive-going pulse "overshoots." Running the 6BE6 controlled i.f. amplifier at low screen voltage makes it possible for the No. 3 grid (pin 7) to cut off the stage at a lower voltage than if the screen were operated at the more-normal 100 volts, but it also reduces the available gain through the stage.

It is necessary to avoid i.f. feedback around the 6BE6 stage, and the closer RFC_1 can be to self-resonant at the i.f. the better will be the filtering. The filtering cannot be improved by increasing the values of the 150-$\mu\mu$f. capacitors because this will tend to "stretch" the pulses and reduce the signal strength when the silencer is operative.

SIGNAL-STRENGTH AND TUNING INDICATORS

The simplest tuning indicator is a milliammeter connected in the d.c. plate lead of an a.g.c.-controlled r.f. or i.f. stage. Since the plate current is reduced as the a.g.c. voltage becomes higher with a stronger signal, the plate current is a measure of the signal strength. The meter can have a 0-1, 0-2 or 0-5 ma. movement, and it should be shunted by a 25-ohm rheostat which is used to set the no-signal reading to full scale on the meter. If a "forward-reading" meter is desired, the meter can be mounted upside down.

Two other S-meter circuits are shown in Fig. 5-23. The system at A uses a milliammeter in a bridge circuit, arranged so that the meter readings increase with the a.v.c. voltage and signal strength. The meter reads approximately in a linear decibel scale and will not be "crowded."

To adjust the system in Fig. 5-23A, pull the tube out of its socket or otherwise break the cathode circuit so that no plate current flows, and adjust the value of resistor R_1 across the meter until the scale reading is maximum. The value of resistance required will depend on the internal resistance of the meter, and must be determined by trial and error (the current is approximately 2.5 ma.). Then replace the tube, allow it to warm up, turn the a.g.c. switch to "off" so the grid is shorted to ground, and adjust the 3000-ohm variable resistor for zere meter current. When the a.g.c. is "on," the meter will follow the signal variations up to the point where the voltage is high enough to cut off the meter tube's plate current. With a 6J5 or 6SN7GT this will occur in the neighborhood of 15 volts, a high-amplitude signal.

The circuit of Fig. 5-23B requires no additional tubes. The resistor R_2 is the normal cathode

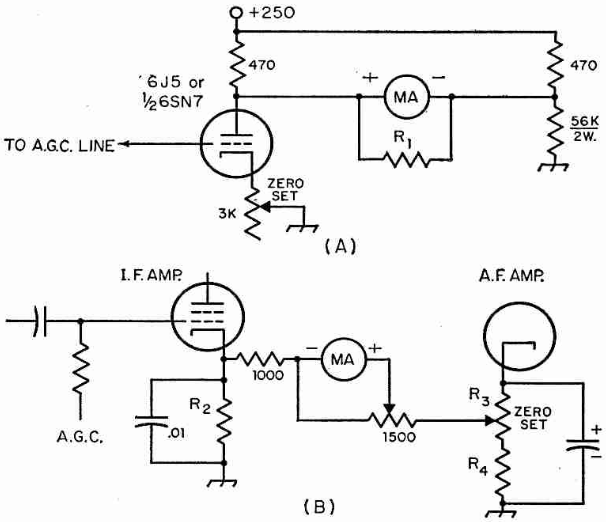


Fig. 5-23—Tuning indicator or S-meter circuits for superheterodyne receivers.

MA—0-1 or 0-2 milliammeter. R_1–R_4—See text.

resistor of an a.g.c.-controlled i.f. stage; its cathode resistor should be returned to chassis and not to the manual gain control. The sum of R_3 plus R_4 should equal the normal cathode resistor for the audio amplifier, and they should be proportioned so that the arm of R_3 can pick off a voltage equal to the normal cathode voltage for the i.f. stage. In some cases it may be necessary to interchange the positions of R_3 and R_4 in the circuit.

The zero-set control R_3 should be set for no reading of the meter with no incoming signal, and the 1500-ohm sensitivity control should be set for a full meter reading with the i.f. tube removed from its socket.

Neither of these S-meter circuits can be "pinned", and only severe misadjustment of the zero-set control can injure the meter.

● HEADPHONES AND LOUDSPEAKERS

There are two basic types of headphones in common use, the magnetic and the crystal. A magnetic headphone uses a small electromagnet that attracts and releases a steel diaphragm in accordance with the electrical output of the radio receiver; this is similar to the "receiver" portion of the household telephone. A crystal headphone uses the piezoelectric properties of a pair of Rochelle-salt or other crystals to vibrate a diaphragm in accordance with the electrical output of the radio receiver. Magnetic headphones can be used in circuits where d.c. is flowing, such as the plate circuit of a vacuum tube, provided the current is not too heavy to be carried by the wire in the coils; the limit is usually a few milliamperes. Crystal headphones can be used only on a.c. (a steady d.c. voltage will damage the crystal unit), and consequently must be coupled to a tube through a device, such as a capacitor or transformer, that isolates the d.c. but passes the a.c. Most modern receivers have a.c. coupling to the headphones and hence either type of headphone can be used, but it is wise to look first at the circuit diagram in the instruction book and make sure that the headphone jack is connected to the secondary of the output transformer, as is usually the case.

In general, crystal headphones will have considerably wider and "flatter" audio response than will magnetic headphones (except those of the "hi-fi" type that sell at premium prices). The lack of wide response in the magnetic headphones is sometimes an advantage in code reception, since the desired signal can be set on the peak and be given a boost in volume over the undesired signals at slightly different frequencies.

Crystal headphones are available only in high-impedance values around 50,000 ohms or so, while magnetic headphones run around 10,000 to 20,000 ohms, although they can be obtained in values as low as 15 ohms. Usually the impedance of a headphone set is unimportant because there is more than enough power available from the radio receiver, but in marginal cases it is possible to improve the acoustic output through a better match of headphone to output impedance. When headphone sets are connected in series or in parallel they must be of similar impedance levels or one set will "hog" most of the power.

Loud speakers are practically always of the low-impedance permanent-field dynamic variety, and the loudspeaker output connections of a receiver can connect directly to the voice coil of the loudspeaker. Some receivers also provide a "500-ohm output" for connection to a long line to a remote loudspeaker. A loudspeaker requires mounting in a suitable enclosure if full low-frequency response is to be obtained.

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 side bands without impairing the quality at all. Maximum receiver selectivity in phone reception requires good 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 100 or 200 *cycles* for hand-key speeds, but this much selectivity requires good stability in both transmitter and receiver, and a slow receiver tuning rate for ease of operation.

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 not obtained with nonregenerative amplifiers using ordinary tuned circuits unless a low i.f. or a large number of circuits is used.

Regeneration

Regeneration can be used to give a 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 feedback 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. However, the regenerative gain varies with signal strength, being less on strong signals.

Crystal-Filters; Phasing

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 crystal is ground resonant at the i.f. and used as a selective coupler between i.f. stages.

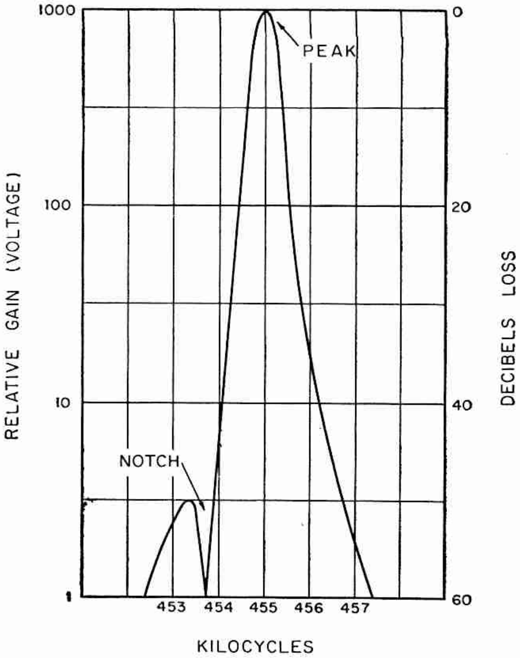


Fig. 5-24—Typical response curve of a crystal filter. The notch can be moved to the other side of the response peak by adjustment of the "phasing" control. With the above curve, setting the b.f.o. at 454 kc. would give good single-signal c.w. reception.

Fig. 5-24 gives a typical crystal-filter resonance curve. For single-signal reception, the audio-frequency image can be reduced by 50 db. or more. Besides practically eliminating the a.f. image, the high selectivity of the crystal filter provides good discrimination against adjacent signals and also reduces the noise.

Two crystal-filter circuits are shown in Fig. 5-25. The circuit at A (or a variation) is found in many of the current communications receivers. The crystal is connected in one side of a bridge circuit, and a ***phasing*** capacitor, C_1, is connected in the other. When C_1 is set to balance the crystal-holder capacitance, the resonance curve of the filter is practically symmetrical; the crystal acts as a series-resonant circuit of very high Q and allows signals over a narrow band of frequencies to pass through to the following tube. More or less capacitance at C_1 introduces the "rejection notch" of Fig. 5-24 (at 453.7 kc. as drawn). The Q of the load circuit for the filter is adjusted by the setting of R_1, which in turn varies the bandwidth of the filter from "sharp" to a bandwidth suitable for phone reception. Some of the components of this filter are special and not generally available to amateurs.

The "band-pass" crystal filter at B uses two crystals separated slightly in frequency to give a band-pass characteristic to the filter. If the frequencies are only a few hundred cycles apart, the characteristic is an excellent one for c.w.

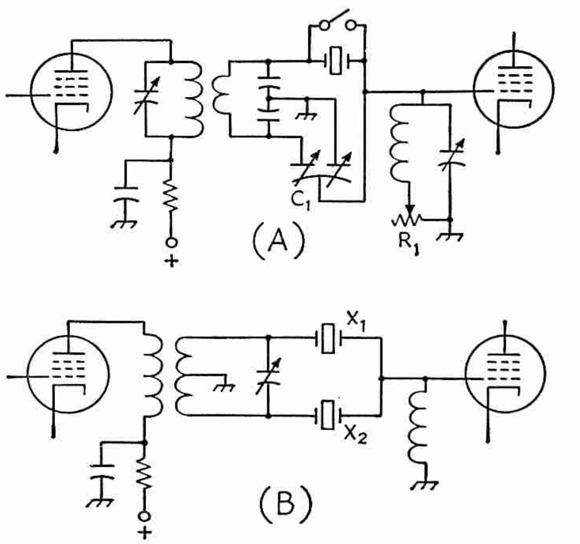


Fig. 5-25—A variable-selectivity crystal filter (A) and a band-pass crystal filter (B).

reception. With crystals about 2 kc. apart, a good phone characteristic is obtained.

Additional I.F. Selectivity

Many commercial communications receivers do not have sufficient selectivity for amateur use, and their performance can be improved by additional i.f. selectivity. One method is to loosely couple a BC-453 aircraft receiver (war surplus, tuning range 190 to 550 kc.) to the tail end of the 455-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 sharp for voice work — 6.5 kc. wide at —60 db. — and it helps considerably in separating phone signals and in backing up crystals filters for improved c.w. reception.

If a BC-453 is not available, one can still enjoy the benefits of improved selectivity. It is only necessary to heterodyne to a lower frequency the 455-kc. signal existing in the receiver i.f. amplifier and then rectify it after passing it through the sharp low-frequency amplifier. The J. W. Miller Company offers 50-kc. transformers for this application.

● 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 a stage of r.f. amplification, for image rejection and a good noise figure (mixers are noisier than amplifiers).

Receivers with an i.f. of 455 kc. can be expected to have some r.f. image response at a signal frequency of 14 Mc. and higher if only one stage of r.f. amplification is used. (Regeneration in the r.f. amplifier will reduce image response, but regeneration usually requires frequent readjustment when tuning across a band.) 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. A common solution at 28 Mc. is to use a "double-conversion" superheterodyne, with one stage of r.f. amplification and a first i.f. of 1600 kc. or higher. A normal receiver with an i.f. of 455 kc. can be converted to a double conversion by connecting a "converter" 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 tuned circuits.

Transistor R. F. Amplifier

A typical r.f. amplifier circuit using a 2N370 transistor is shown in Fig. 5-26. Since it is desirable to maintain a reasonable Q in the tuned circuits, to reduce r.f. image response, the base and collector are both tapped down on their tuned circuits. An alternative method, using low-impedance inductive coupling, is shown in Fig. 5-26B; this method is sometimes easier to adjust than the taps illustrated in Fig. 5-26A. The tuned

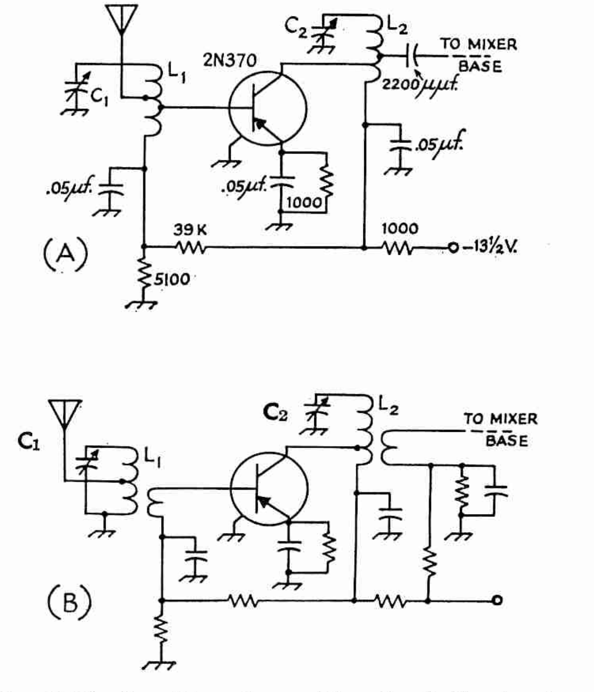


Fig. 5-26—Transistor r.f. amplifier circuit. The low-impedance connections to the base and collector can be (A) taps on the inductors or (B) low-impedance coupling links. L_1C_1, L_2C_2—Resonant at signal frequency.

circuits, L_1C_1 and L_2C_2, should resonate at the operating frequency, and they should be mounted or shielded to eliminate inductive coupling between each other.

FEEDBACK

Feedback giving rise to regeneration and oscillation can occur in a single stage or it may appear as an over-all feedback through several stages that are on the same frequency. To avoid feedback 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. 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 means good shielding of coils and tuning capacitors in r.f. and i.f. circuits, the use of good bypass capacitors (mica or ceramic at r.f., paper or ceramic at i.f.), and returning all bypass capacitors (grid, cathode, plate and screen) for a given stage with short leads to one spot on the chassis. If single-ended tubes are used, the screen or cathode bypass capacitor 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 feedback in a multistage amplifier, 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 bypassing 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 bandwidth 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 working at high gain. It shows up 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. The 6BJ6, 6BA6 and 6DC6 are recommended for r.f. amplifiers where cross-modulation may be a problem.

A receiver designed for minimum cross-modulation will use as little gain as possible ahead of the high-selectivity stages, to hold strong unwanted signals below the overload point.

Gain Control

To avoid cross-modulation and other overload effects in the mixer 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.g.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 schematic form in Fig. 5-27.

Tracking

In a receiver with no r.f. stage, it is no incon-

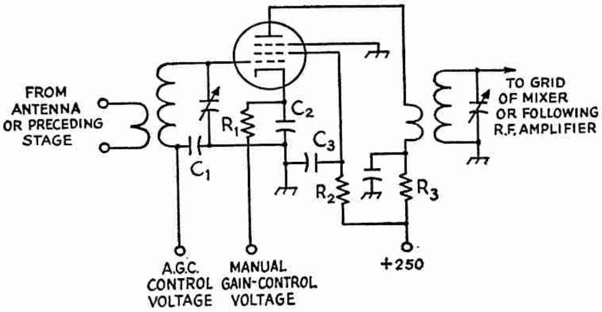


Fig. 5-27—Typical radio-frequency amplifier circuit for a superheterodyne receiver. Representative values for components are as follows:

C_1 to C_4—0.01 μf. below 15 Mc., 0.001 μf. at 30 Mc.
R_1, R_2—See Table 5-II.
R_3—1800 ohms.

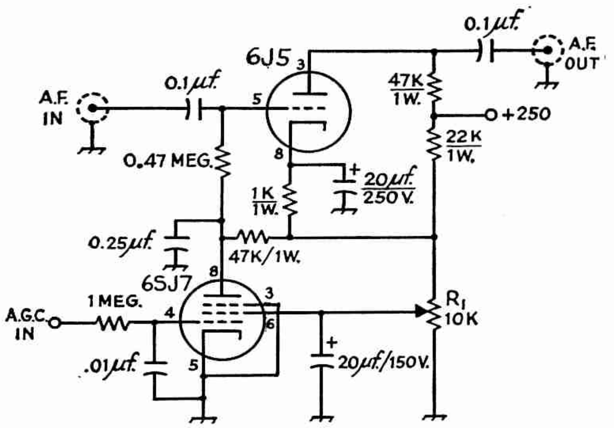


Fig. 5-28 — A practical squelch circuit for cutting off the receiver output when no signal is present.

venience 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 r.f. stages and mixer will require retuning over an entire amateur band. 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.**

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. 5-9A, 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 (semicircular plates).

Squelch Circuits

An audio squelch circuit is one that cuts off the receiver output when no signal is coming through the receiver. It is useful in mobile or net work where the no-signal receiver noise may be as loud as the signal, causing undue operator fatigue during no-signal periods.

A practical squelch circuit is shown in Fig. 5-28, When the a.g.c. voltage is low or zero, the 6SJ7 draws plate current. Voltage drop across the 47,000-ohm resistor in its plate circuit cuts off the 6J5 and no receiver signal or noise is passed. When the a.g.c. voltage rises to the cut-off value of the 6SJ7, the pentode no longer draws current and the bias on the 6J5 is now only the operating bias, furnished by the 1000-ohm cathode resistor. The triode now functions as an ordinary amplifier and passes signals. By varying the screen voltage on the 6SJ7 through R_1, the pentode's cut-off bias can be varied, so that the relation between a.g.c. voltage and signal cut-off point of the amplifier is adjustable.

Connections to the receiver consist of two a.f. lines (shielded), the a.g.c. lead, and chassis ground. The squelch circuit is normally inserted between detector output and the audio volume control of the receiver. Since the circuit is used in the low-level audio point, its plate supply must be free from a.c. or objectionable hum will be introduced.

Improving Receiver Sensitivity

The sensitivity (signal-to-noise ratio) of a receiver on the higher frequencies above 20 Mc. is dependent upon the band width of the receiver and the noise contributed by the "front end" of the receiver. Neglecting the fact that image rejection may be 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 than 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, 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 cannot give best selectivity in the antenna circuit, so it is futile to try to maximize sensitivity *and* selectivity in this circuit.

When a receiver is satisfactory in every respect (stability and selectivity) except sensitivity on 14 through 30 Mc., the best solution for the amateur is to add a **preamplifier,** a stage 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. This can be accomplished by changing the antenna feed line 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.

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 amateurs have used it with considerable success. High-g_m tubes are the best as regenerative amplifiers, and the feedback 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 feedback coupling. This is a tricky process 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. A good receiver might well have two gain controls, one for the first radio-frequency stage and another for the i.f. and other r.f. stages.

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 best receiver condition for the reception of code signals will have the first r.f. stage running at maximum gain, the following r.f., mixer and i.f. stages operating with just enough gain to maintain the signal-to-noise ratio, and the audio gain set to give comfortable headphone or speaker volume. The audio volume should be controlled by the audio gain control, not the i.f. gain control. Under the above conditions, the selectivity of the receiver is being used to best advantage, and cross-modulation is minimized. It precludes the use of a receiver in which the gains of the r.f. and i.f. stages are controlled simultaneously.

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 set at its sharpest position, if variable selectivity is available. The initial adjustment should be made with the phasing control in an intermediate position. Once adjusted, the beat oscillator should be left set and the receiver tuned to the other side of zero beat (audio-frequency image) on the same signal 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 other.

An interfering signal having a beat note differing from that of the a.f. image can be similarly phased out, provided its frequency is not too near the desired signal.

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 high selectivity often makes it difficult to find the desired station quickly, if the filter is switched in only when interference is present.

A.M. Phone Reception

In reception of a.m. phone signals, the normal procedure is to set the r.f. and i.f. gain at maximum, switch on the a.g.c., and use the audio gain control for setting the volume. This insures maximum effectiveness of the a.g.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.g.c., in which case the weaker station may disappear because of the reduced gain. In this case better reception may result if the a.g.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.

When receiving an a.m. signal on a frequency within 5 to 20 kc. from a single-sideband signal it may also be necessary to switch off the a.g.c. and resort to the use of manual gain control, unless the receiver has excellent skirt selectivity. No ordinary a.g.c. circuit can handle the syllabic bursts of energy from the sideband station, but there are special circuits that will.

A crystal filter will help reduce interference in phone reception. Although the high selectivity cuts sidebands and reduces the audio output at the higher audio frequencies, it is possible to use quite high selectivity without destroying intelligibility. As in code 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.

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 circuits, reduces 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 desired signals peak.

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.

Alignment and Servicing of Superheterodyne Receivers

I.F. Alignment

A calibrated signal generator or test oscillator is a useful device for 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. Lacking an S meter, a high-resistance voltmeter or 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-μf. blocking capacitor 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.g.c. of the receiver should be turned on, but any other indication requires that it be turned off. Lacking a test oscillator, a steady signal 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 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 capacitor to the grid of the last i.f. amplifier tube. The trimmer capacitors 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. It is desirable in all cases to use the minimum 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 the receiver will also be described in detail.

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 capacitor 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 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 capacitor) 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 capacitor, 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 capacitor 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, more inductance is needed; conversely, 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 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 shows up as 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.g.c. system to reduce the receiver gain drastically. Oscillation can be caused by poor connections in the common ground circuits. Inadequate or defective bypass capacitors 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. Inadequate screen or plate bypass capacitance is a common cause of such oscillation.

Improving the Performance of Receivers

Frequently amateurs unjustly criticize a receiver's performance when actually part of the trouble lies with the operator, in his lack of knowledge about the receiver's operation or in his inability to recognize a readily curable fault. The best example of this is a complaint about "lack of selectivity" when the receiver contains an i.f. crystal filter and the operator hasn't bothered to learn how to use it properly. "Lack of sensitivity" may be nothing more than poor alignment of the r.f. and mixer tuning. The cures for these two complaints are obvious, and the details are treated both in this chapter and in the receiver instruction book.

However, many complaints about selectivity, sensitivity, and other points are justified. Inexpensive, and most second-hand, receivers cannot be expected to measure up to the performance standards of some of the current and top-priced receivers. Nevertheless, many amateurs overlook the possibility of improving the performance of these "bargains" (they may or may not be bargains) by a few simple additions or modifications. From time to time articles in *QST* describe improvements for specific receivers, and it may repay the owner of a newly-acquired second-hand receiver to examine past issues and see if an applicable article was published. The annual index in each December issue is a help in this respect.

Where no applicable article can be found, a few general principles can be laid down. If the complaint is the inability to separate stations, better i.f. (and occasionally audio) selectivity is indicated. The answer is not to be found in better bandspread tuning of the dial as is sometimes erroneously concluded. For code reception the addition of a "*Q* Multiplier" to the i.f. amplifier is a simple and effective attack; a *Q* Multiplier is at its best in the region 100 to 900 kc., and higher than this its effectiveness drops off. The Selectoject is a selective audio device based on similar principles. For phone reception the addition of a *Q* Multiplier will help to reject an interfering carrier, and the use of a BC-453 as a "Q5-er" will add adjacent-channel selectivity.

With the addition of more i.f. selectivity, it may be found that the receiver's tuning rate (number of kc. tuned per dial revolution) is too high, and consequently the tuning with good i.f. selectivity becomes too critical. If this is the case, a 5-to-1 reduction planetary dial drive mechanism may be added to make the tuning rate more favorable. These drives are sold by the larger supply houses and can usually be added to the receiver if a suitable mounting bracket is made from sheet metal. If there is already some backlash in the dial mechanism, the addition of the planetary drive will magnify its effect, so it is necessary to minimize the backlash before attempting to improve the tuning rate. While this is not possible in all cases, it should be investigated from every angle before giving up. Replacing a small tuning knob with a larger one will add to ease of tuning; in many cases after doing so it will then be desirable or necessary to raise the receiver higher above the table.

If the receiver appears to lack the ability to bring in the weak signals, particularly on the higher-frequency bands, the performance can often be improved by the addition of an antenna coupler (described elsewhere in this chapter); it will always be improved by the addition of a preselector (also described elsewhere in this chapter).

If the receiver shortcoming is inadequate r.f. selectivity, as indicated by r.f. "images" on the higher-frequency bands, a simple antenna coupler will often add sufficient selectivity to cure the trouble. However, if the images are severe, it is likely that a preselector will be required, preferably of the regenerative type. The preselector will also add to the ability of the receiver to detect weak signals at 14 Mc. and higher.

In many of the inexpensive receivers the frequency calibration of the dial is not very accurate. The receiver's usefulness for determining band limits will be greatly improved by the addition of a 100-kc. crystal-controlled frequency standard. These units can be built or purchased complete at very reasonable prices, and no amateur station worthy of the name should be without one.

Some receivers that show a considerable frequency drift as they are warming up can be improved by the simple expedient of furnishing more ventilation, by propping up the lid or by drilling extra ventilation holes. In many cases the warm-up drift can be cut in half.

Receivers that show frequency changes with line-voltage or gain-control variations can be greatly improved by the addition of regulated voltage on the oscillators (high-frequency and b.f.o.) and the screen of the mixer tube. There is usually room in any receiver for the addition of a VR tube of the right rating.

SimpleX Super

The "SimpleX Super" Three-Tube Receiver

The name of this receiver derives from "simple", "X" for crystal (filter), and "super" for superheterodyne; hence a "simple crystal-filter superheterodyne." For about fifty dollars and a few nights at the workbench this little receiver will allow you to copy practically any c.w. or s.s.b. signal in the 40- or 80-meter band that a much more expensive receiver might drag in. By the flip of a switch you can tune to 5 Mc. for WWV.

This 3-tube receiver will permit the single-signal reception of code signals. Single-sideband phone can be handled with no difficulty at all. With the b.f.o. turned off for the reception of a.m. signals, a threshold effect shows up that prevents digging all the way down for the weak ones, but one can still copy plenty of a.m. signals. Since the receiver uses only three tubes, it doesn't have the more-than-enough gain of a big receiver, and its performance won't be very impressive on a poor (short or low) antenna. However, if the transmitting antenna is also used for receiving, you will find yourself backing down on the volume control to save your ears.

Referring to the circuit diagram in Fig. 5-30, the receiver is a superheterodyne with an intermediate frequency of 1700 kc. With the h.f. oscillator tuning 5.2 to 5.7 Mc., the 3.5- or the 7-Mc. amateur bands can be tuned merely by retuning the input circuit, L_1C_1. Since C_1 is large enough to hit the two bands without a coil change, the band-changing process consists of turning C_1 to the low- or high-capacitance end of its range. To copy WWV at 5 Mc., the oscillator must be tuned to 3.3 Mc., and this is done by switching in an additional capacitor across the oscillator circuit.

If you are disappointed because the receiver doesn't tune the 21-Mc. band, remember that the "under-$100" receivers don't either. Sure, the dials show 21 Mc., but try to use the receivers to hold a signal for any length of time! The SimpleX Super, with a crystal-controlled converter between it and the antenna, will handle 15 meters like 80.

Selectivity at the i.f. is obtained through the use of a single crystal. Although not as sharp as the usual 455-kc. crystal filter, it is sharp enough to provide a fair degree of single-signal c.w. reception and yet broad enough for good copy of an s.s.b. phone signal.

In the detector stage, the pentode section of a 6U8A is used as a grid-leak detector, and the triode section serves as the b.f.o. Stray coupling at the socket and in the tube provides adequate injection. Audio amplification is obtained from the two triode sections of a 6CG7. The primary of a small output transformer, T_1, serves as the coupling for high-impedance headphone output, and a small loudspeaker or low-impedance headphones can be connected at the output winding of the transformer. Although the audio power output is less than a watt, it is sufficient to drive a loudspeaker adequately in a small quiet room.

The power supply uses a large choke and two 40-μf. capacitors, and the very slight hum that can be detected in the headphones with the volume full on is stray a.c. picked up by the detector grid; it doesn't come from inadequate filtering of the power supply. (The hum can only be heard with no antenna on; under normal operation the incoming noise will mask the slight hum.)

A switch at the input of the receiver is included so that the receiver can be used to listen to one's own transmitter without too severe blocking. Using the b.f.o. switch to cut in the WWV padder was done (instead of by the more logical S_1) to keep the input short-circuiting leads short.

An 8 × 12 × 3-inch aluminum chassis takes all of the parts without crowding, and the location of the components can be seen in the photographs. The $7\frac{1}{4} \times 13$-inch aluminum panel ($\frac{1}{16}$-inch thick) is held to the chassis by the b.f.o. capacitor mounting screws, the phone jack, the dial drive and the two rotary switches. The tuning capacitor C_2 is mounted on a small aluminum bracket made from an extra strip of the panel material; before the bracket is finally fastened to the chassis the capacitor and bracket should be used to locate the dial hole on the panel. When

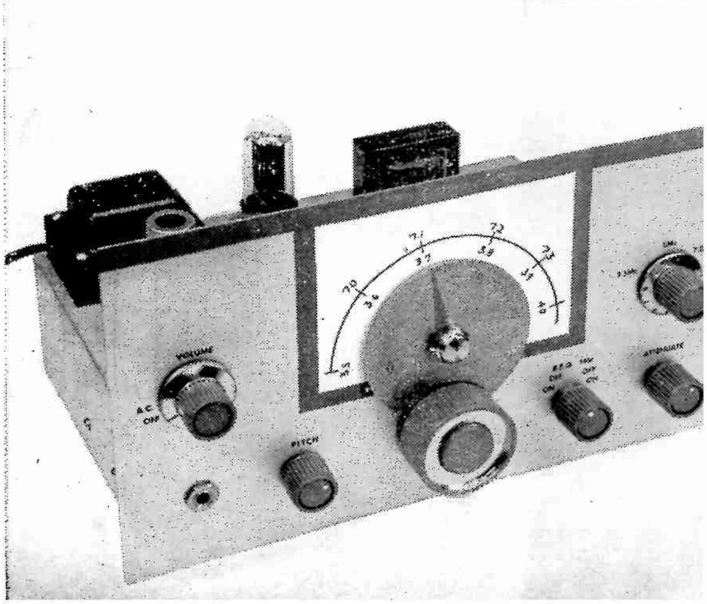


Fig. 5-29—The SimpleX Super receiver uses three dual tubes and a crystal filter to cover the 80- and 40-meter bands, and it can tune to 5 Mc. for copying WWV. The dial scale is made from white paper held to the panel by red Scotch tape; the pointer is a slice of the tape.

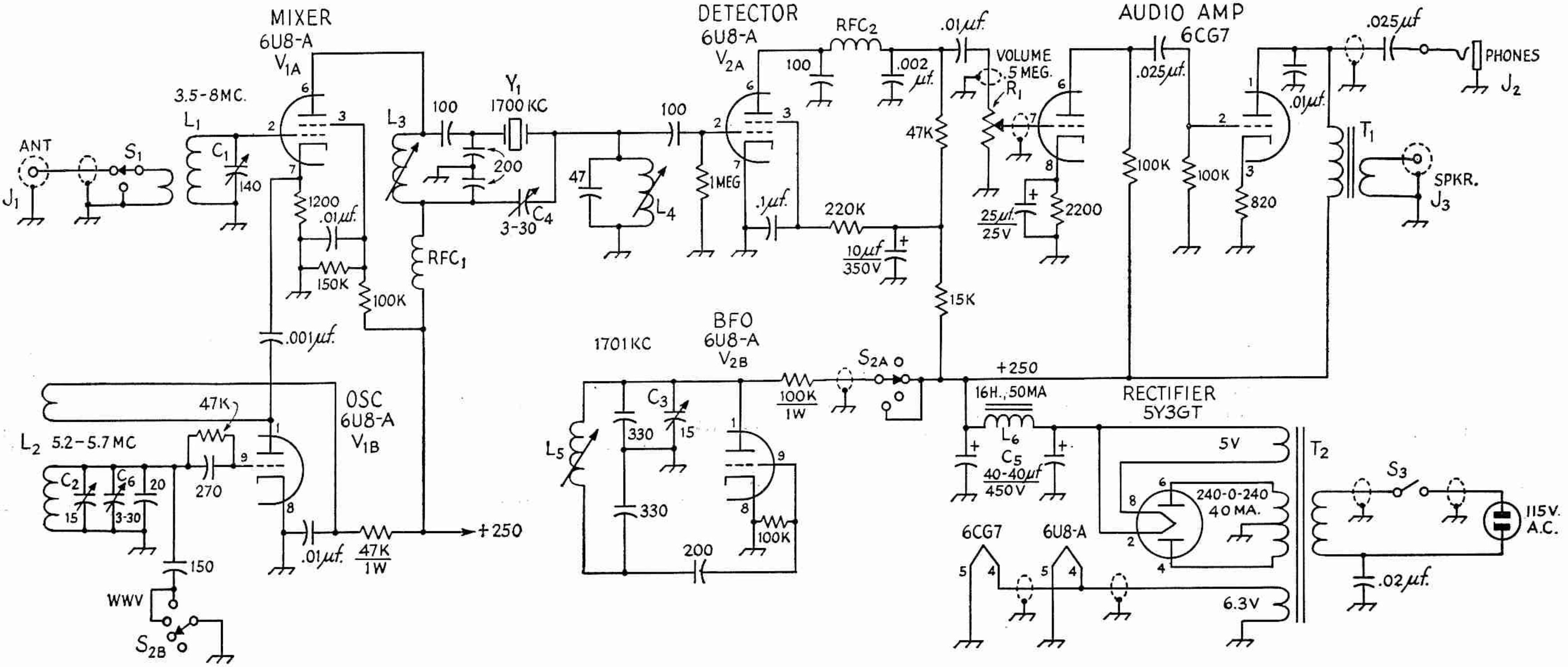


Fig. 5-30—Circuit diagram of the SimpleX Super receiver. Unless otherwise indicated, capacitances are in $\mu\mu$f., resistances are in ohms, resistors are ½ watt. Polarity shown on electrolytic capacitors; fixed capacitors 330 $\mu\mu$f. or less are silver mica or NPO ceramic. Nonelectrolytic fixed capacitors over 0.025 μf. are 400-volt molded tubulars. Fixed capacitors 0.001 through 0.025 are ceramic.

C_1—140-$\mu\mu$f. midget variable (Hammarlund APC-140-B).
C_2—15-$\mu\mu$f. midget variable (Hammarlund HF-15).
C_3—15-$\mu\mu$f. trimmer (Hammarlund MAPC-15-B).
C_4, C_6—3–30-$\mu\mu$f. mica compression trimmer.
C_5—Dual 40-μf. 450-volt electrolytic (Mallory TCD-78 or equiv.).
J_1, J_3—Phono jack.
J_2—Open-circuit headphone jack.
L_1, L_2—See Fig. 5-35.
L_3, L_4—105–200-μh. slug-tuned (North Hills 120-H coil mounted in North Hills S-120 shield can).
L_5—36–64-μh. slug-tuned (North Hills 120-F coil mounted in North Hills S-120 shield can).
L_6—16-hy. 50-ma. filter choke (Knight 62-G-137 or equiv.).
R_1—½ megohm volume control, audio taper, with switch.
RFC_1, RFC_2—2.5-mh. r.f. choke (Waters C1155).
S_1—1-pole 12-position (2 used) rotary ceramic switch (Centralab PA-2001).
S_2—2-pole 6-position (4 used) rotary ceramic switch (Centralab PA-2003).
S_3—S.p.s.t. switch, part of R_1.
T_1—10,000-ohms-to-voice-coil output transformer (Stancor A-3822 or equiv.).
T_2—480 v. c.t. at 40 ma., 5 v. at 2 amp., 6.3 v. at 2 amp. (Knight 62-G-034 or equiv.).
Y_1—1700-kc. crystal in FT-243 holder (E. B. Lewis or equiv.).

(All radio stores do not handle the above components. For prices and names of dealers write to North Hills Electric Co., 402 Sagamore Ave., Mineola, N. Y.; Knight is handled by Allied Radio, 100 N. Western Ave., Chicago 80, Ill.; Waters Mfg. Inc., Boston Post Rd., Wayland, Mass.; E. B. Lewis, 11 Bragg St., E. Hartford, Conn.)

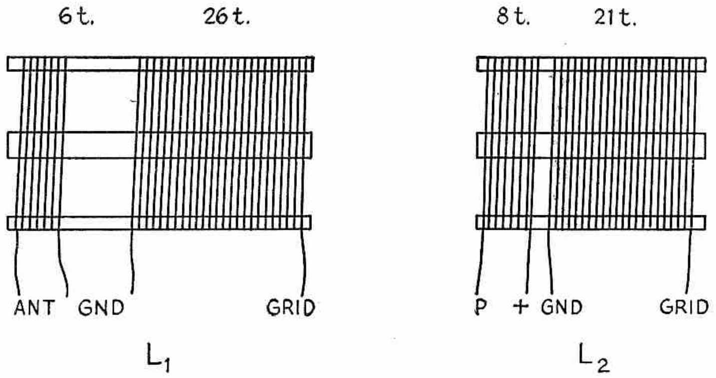


Fig. 5-31—Details of the coil construction. Each one is made from B & W 3016 Miniductor stock, which is wound 32 t.p.i. and 1-inch diameter. The separation between coils in L_1 is 7 turns; the separation between coils L_2 is 1 turn. It is important that the coils be connected as indicated. The Miniductor stock can be cut into the required lengths by pushing in a turn, cutting it inside the coil and then pushing the newly cut ends through to outside the coil. Once outside, it is easy to peel away the wire with the help of long-nose pliers. When sufficient turns have been removed, the support bars can be cut with a fine saw.

drilling the hole for the dial drive, measure the dimension instead of using the template provided with the National K dial. It pays to take care in mounting the tuning capacitor and the dial, since a smooth tuning drive is an essential in any receiver. To facilitate tuning, a National HRT knob was used instead of the puny knob furnished with the K dial. The other knobs are gray National HR and HR-4.

Tie points are used liberally throughout the receiver, as junctions for components and interconnecting wires. The coils L_1 and L_2 are mounted on tie points, using short leads. If the leads from L_2 are too long, the coil will be "floppy" and the receiver may be unstable. Fig. 5-35 shows how coils L_1 and L_2 are constructed and connected. The leads from C_1 and C_2 are brought through the chassis in insulating grommets. The 3- to 30-$\mu\mu$f. mica compression trimmer across L_2 is soldered to the tie points that support the coil.

The receiver is wired with shielded wire for many of the leads, in an effort to minimize hum in the audio and feedthrough around the crystal filter. The shielded leads are marked in Fig. 5-30 where feasible; the simple rule to follow is to shield all B+ leads along with those shown shielded in Fig. 5-30. For easy of wiring, these shielded leads should be installed first or at least early in the construction. As the wiring progresses, a neat-looking unit can be obtained by dressing the leads and components in parallel lines or at right angles. D.c. and a.c. leads can be tucked out of the way along the edges of the chassis, while r.f. leads should be as direct as is reasonable.

If this is your first receiver or construction job, there are several pitfalls to be avoided. When installing a tube socket, first give a little thought to where the grid and plate leads will run, and orient the socket so that these leads will be direct and not cross over the socket.

Another thing to look out for is the well-meaning store clerk who sells you stranded wire for making the connections throughout the receiver. The only stranded wire in this receiver is in the leads from the transformers, filter capacitor and filter choke, and in the shielded wire, and all this only because there was no choice. Where stranded wire is used, be very careful to avoid wild strands that stray over to an adjacent socket terminal and short-circuit a part of the circuit without your knowing it. No. 20 or 22 insulated solid tinned copper wire should be used for connections wherever no shielding is used. Long bare leads from resistors or capacitors should be covered with insulating tubing unless they go to chassis grounds.

The final bugaboo is, of course, a poorly-soldered connection. If this is your first venture, by all means practice soldering before you start to wire this receiver. Read an article or two on how to solder, or get a friend to show you how and to criticize your first attempts. A good soldering iron is an essential; there have been instances of a first venture having been "soldered" with an iron that would just barely melt the solder; the iron was incapable of heating the solder and work to the point where the solder would flow properly.

There is no need to worry about the dial scale when the receiver is first built, because the receiver has to be checked. The scale is a sheet of white paper held in place by red or black Scotch tape. The dial pointer is a slice of the same tape.

When the wiring has been completed and checked once more against the circuit diagram, plug in the tubes and the line cord and turn on the receiver through S_3. The tube heaters and rectifier filament should light up and nothing should start to smoke or get hot. If you have a voltmeter you should measure about 250 volts on the B+ line.

With headphones plugged in the receiver, you should be able to hear a little hum when the volume control is advanced all the way. If you can't hear any hum, touching a screwdriver to Pin 2 should produce hum and a loud click. This shows that the detector and audio amplifier are working.

The next step is to tune L_3, L_4 and L_5 to 1700 kc., the crystal frequency. If you have or can borrow a signal generator, put 1700-kc. r.f. in at the grid of the 6U8A mixer and peak L_3 and L_4. Lacking a signal generator, you may be lucky enough to find a strong signal by tuning around with C_2, but it isn't likely. Your best bet is to tune a broadcast receiver to around 1245 kc.; if the receiver has a 455-kc. i.f. the oscillator will then be on 1700 kc. Don't depend upon the calibration of the broadcast receiver; make your own by checking known stations. The oscillator of the broadcast receiver will furnish a steady (possibly hum-modulated) carrier that can be picked up by running a wire temporarily from the grid of the 6U8A mixer to a point near the chassis of the

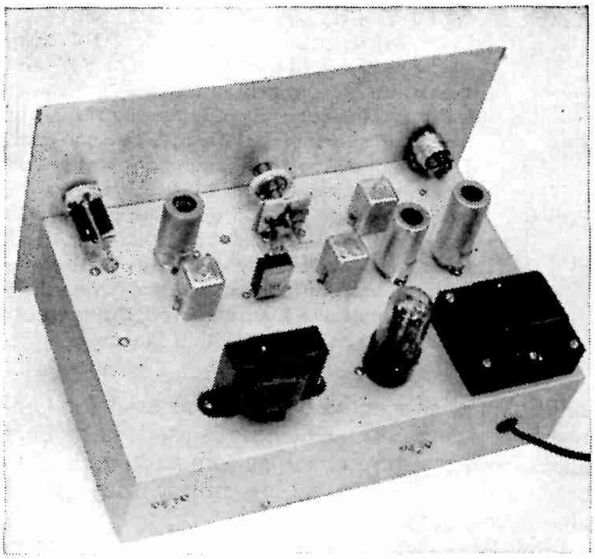


Fig. 5-32—Top view of the SimpleX Super. The tube between the two variable capacitors is the mixer-oscillator 6U8A; the 6CG7 audio amplifier is at the far right. The flexible insulated coupling between main tuning dial and the tuning capacitor is a Millen 39016.

b.c. receiver. Adjust L_5 until you get a beat with the 1700-kc. signal, and then peak L_3 and L_4. If the signal gets too loud, reduce the signal by moving the wire away from the b.c. receiver. Now slowly swing the signal frequency back and forth with the b.f.o. turned off; you should find a spot where the noise rushes up quickly and then drops off. This is the crystal frequency, and L_3 and L_4 should be peaked again on this frequency if you were a little off the first time.

An antenna connected to the receiver should now permit the reception of signals. With C_1 nearly unmeshed, you will be in the region of the 7-Mc. band, and with C_1 almost completely meshed, you will be near 3.5 Mc. Do your tuning with the compression trimmer in the oscillator circuit, until you find a known frequency (it can be your own transmitter). Let's say your transmitter has a crystal at 3725 kc. Set C_2 at half capacitance and tune with C_6 until you hear your transmitter. You shouldn't need any antenna on the receiver for this test. Once you have the setting for the trimmer, put the antenna on the receiver and look around for other known signals. (CHU, the Canadian standard-frequency station at 7335 kc., is a good marker.) With luck you should just be able to cover the 80-meter band; if you can get one end but not the other, a minor readjustment of the trimmer is indicated.

Once you have acquainted yourself with the 80- and 40-meter bands, and appreciate that you have to peak up the input circuit (C_1) fairly often as you tune across the bands, you are ready to trim up the crystal filter. Run the volume fairly high, so that you can hear noise from the properly peaked input circuit, and turn C_3 until the noise takes on a higher-pitched characteristic. (The b.f.o. stage is originally set up with C_3 at midcapacitance and L_5 adjusted for lowest-pitched noise.) Now tune in a code signal with C_2 and swing back and forth through it. "One side" of the signal should be louder than the other. Tune to the weak side with a beat note of around 800 cycles and then adjust C_4 for minimum signal. After a few attempts, juggling C_3, C_4, L_3 and L_4, you should get a condition where the single-signal c.w. effect is quite apparent.

All that remains is to install the dial scale and calibrate it. A 100-kc. oscillator is ideal for this job; lacking one or the ability to borrow one, you will have to rely on other signals. If your crystal filter is 1700 kc. exactly, the 80- and 40-meter calibrations will coincide as they do on the scale shown in Fig. 5-33; if not, the calibration marks will be offset on the two bands.

If you find that you can't get WWV at 5 Mc. with the 150-$\mu\mu$f. capacitor switched in, substitute a 130-$\mu\mu$f. mica in parallel with a 30-$\mu\mu$f. trimmer, and adjust so that WWV falls on scale.

As you acquaint yourself with the operation of the receiver, you will notice that tuning C_1 will have a slight effect on the tuning of the signal. In other words, tuning C_1 "pulls" the oscillator slightly. To remedy this would have made the receiver more complicated, and the simple solution is merely to first peak C_1 on noise and then tune with C_2.

You will find this to be a practical receiver in every way for the c.w. (or s.s.b.) operator. The tuning rate is always the same on 80 or 40, or 15 with a converter, and 21-Mc. s.s.b. signals tune as easily as those on 3.9 Mc. The warm-up drift is negligible, and the oscillator is surprisingly insensitive to voltage changes. Whetber or not the oscillator is insensitive to shock and vibration will depend upon the care with which the components are anchored to their respective tie points.

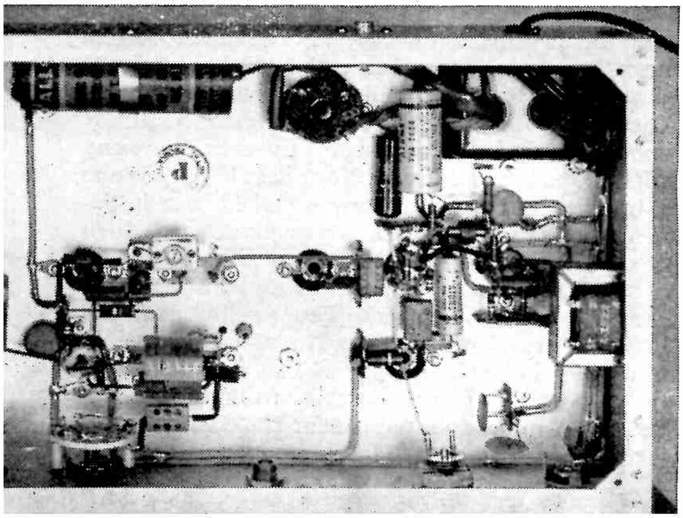


Fig. 5-33—Shielded wire, used for most of the d.c. and 60-cycle leads, lends to the clean appearance underneath the chassis. The switch at the left shorts the input of the receiver, and the adjacent switch handles the b.f.o. and the padding capacitor for WWV.

The phono jack at the top left is for the antenna; the other phono jack is for low-impedance audio output. The headphone jack (lower right) is for high-impedance audio output.

The 2X4+1 Superheterodyne

The receiver shown in Figs. 5-34, 5-37 and 5-38 is a two-band four-tube (2X4) receiver with a transistor (+ 1) 100-kc. frequency standard. Other features include the ability to tune to 5 Mc., for the reception of WWV, and a dual-crystal filter for single-sideband and single-signal c.w. reception. Tuning the 40- and 80-meter amateur bands with good stability and selectivity, the receiver can be used on other bands by the addition of crystal-controlled converters ahead of it.

Referring to the circuit in Fig. 5-35, the pentode section of a 6U8-A is used as a mixer, with the triode portion of the same tube serving as the oscillator. The i.f. is 1700 kc. and the oscillator tunes 5.2 to 5.7 Mc.; tuning the input circuit to the 80-meter band brings in 80-meter signals, and all that is required to hear 40-meter signals is to swing the input tuning, C_1, to the low-capacitance end of its range. Although, e.g., a 7.05-Mc. (5.35 + 1.7) and a 3.65-Mc. (5.35 − 1.7) signal will appear at the same setting of the tuning dial, the two signals cannot be received simultaneously because the double-tuned circuit, C_{1A} L_2 and C_{1B} L_3, between antenna and mixer grid provides the necessary rejection. To provide optimum coupling in both ranges, the coupling capacitance is changed by a switch, S_1, actuated by the shaft of C_1. Thus the coupling change takes place automatically as the capacitor is tuned to the desired band. To make the two circuits track over the entire range, a 3- to 30-$\mu\mu$f trimmer is provided to compensate for the input capacitance of the mixer. For WWV reception, capacitance C_6 is added to the oscillator circuit to bring its frequency to 3.3 Mc.

The mixer is followed by the dual crystal filter at 1700 kc. and a stage of i.f. amplification. I.f. gain is manually controlled by a variable bias control in the cathode circuit of the 6BA6 i.f. amplifier stage. A triode section of a 6CG7, V_{2A}, serves as a grid-leak detector, and the other section is used as the b.f.o. A two-stage audio amplifier follows, providing high-impedance output for headphones or low-impedance output for a loudspeaker. The audio power is sufficient to give more than enough high-impedance headphone volume and quite adequate loudspeaker volume in a quiet room.

The power supply includes a 0C3 to supply regulated 105 volts for the two oscillators and the screen of the mixer.

The transistor 100-kc. calibration oscillator uses for its power source the 8 volts developed across the cathode resistor of V_{3B}. Switch S_3 turns the oscillator on and off and also adds the capacitance to the oscillator circuit that permits WWV reception. The four positions of S_3 are OFF — WWV (only) — CAL (oscillator only) — BOTH. Although the 100-kc. standard is not essential to the operation of the receiver, its inclusion will be found to be quite valuable.

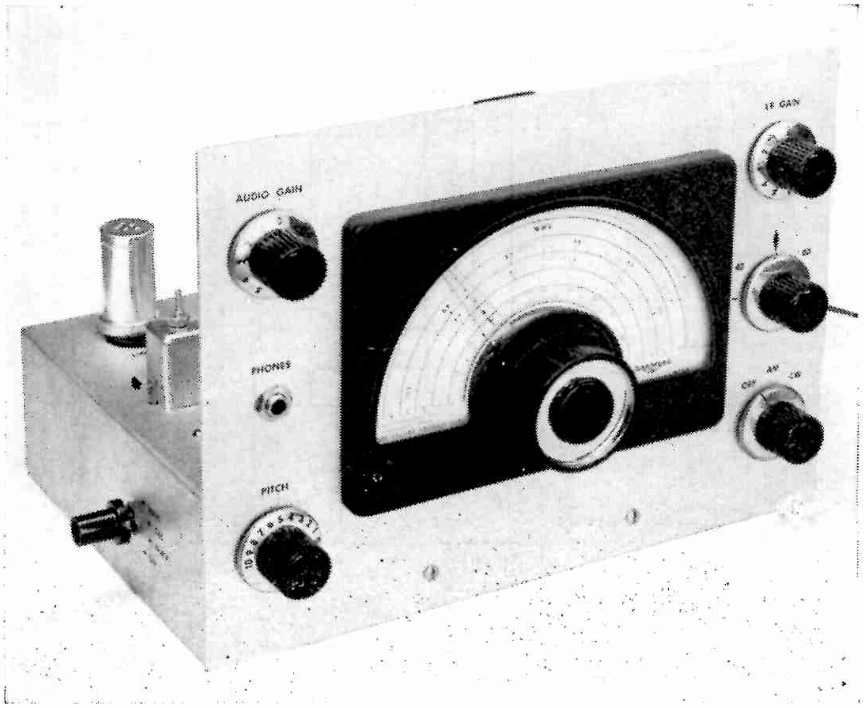


Fig. 5-34—The 2X4 + 1 superheterodyne is a four-tube receiver with 7-tube performance. It tunes the 80- and 40-meter amateur bands, and provision is included for receiving WWV on 5 Mc. A built-in crystal oscillator provides 100-kc. frequency markers throughout the bands. Black knob on the left-hand side controls the calibration oscillator and the WWV reception.

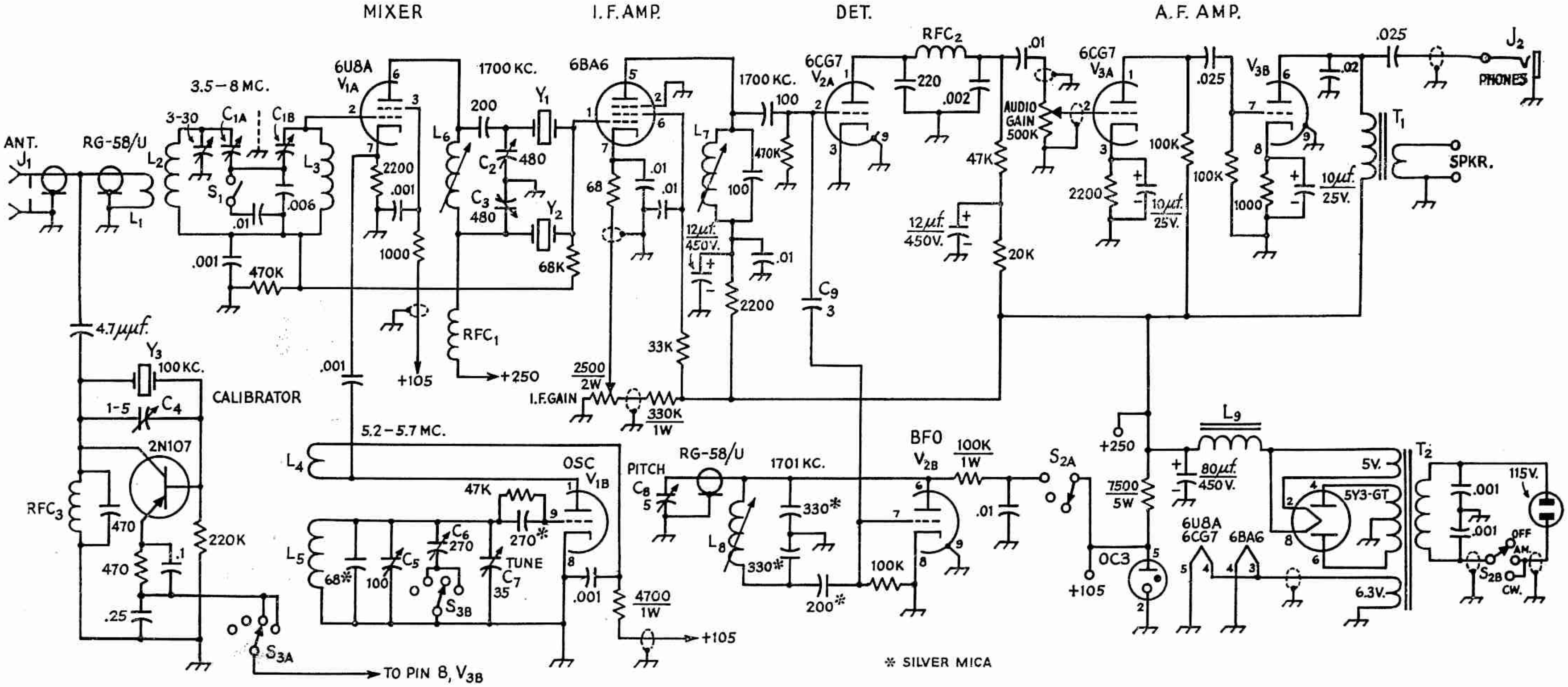


Fig. 5-35—Circuit diagram of the 2X4 + 1 super-heterodyne. Unless indicated otherwise, decimal capacitances are in μf., other capacitances in μμf., resistors are ½ watt.

C_1—Dual variable, 140 μμf. per section (Hammarlund MCD-140-M).

C_2, C_3—480-μμf. mica compression trimmer (Arco-Elmenco 466).

C_4—5-μμf. variable (Hammarlund MAC-5).

C_5—100-μμf. midget variable (Hammarlund HF-100).

C_6—240 μμf. ± 5% mica in parallel with 30-μμf. mica compression trimmer.

C_7—35-μμf. midget variable (Hammarlund HF-35).

C_8—5-μμf. midget variable (Hammarlund HF-15 with 3 plates removed).

C_9—3 μμf. approx. Insulated wires twisted together for 3 turns.

J_1—Phono jack.

L_1—19 turns No. 24, part of L_2 stock, 1/16 inch from L_2.

L_2, L_3—43 turns No. 24, ¾-inch diam, 32 t.p.i. (B&W 3012 or Illumitronic 632).

L_4—7 turns No. 24, part of L_5 stock, 1/32 inch from L_5.

L_5—17 turns No. 24, ¾-inch diam., 32 t.p.i. (B&W 3012 or Illumitronic 632).

L_6, L_7—64 to 105 μh., adjustable (North Hills 120-G in North Hills S-120 shield can).

L_8—36 to 64 μh., adjustable (North Hills 120-F in North Hills S-120 shield can).

L_9—15-henry, 75-ma. filter choke (Stancor C-1002).

RFC_1, RFC_2—2.2 mh., self resonant at 1.6 Mc. (Waters C1059).

RFC_3—10 mh. (National R-50-I).

S_1—Homemade cam switch mounted on C_1. See text.

S_2—2-pole 3-position rotary switch (Centralab 1472).

S_3—2-pole 6-pos tion (4 used) miniature ceramic switch (Centralab PA-3 with Centralab PA-301 index, 2½ inches used).

T_1—3-watt, 8000 to 3.2 ohms, output transformer (Stancor A-3329).

T_2—650 v.c.t. at 55 ma., 5 v., 6.3 v. at 2 amp. (Stancor PC-8407).

Y_1, Y_2—1700-kc. crystals, FT-243 holders, surplus.

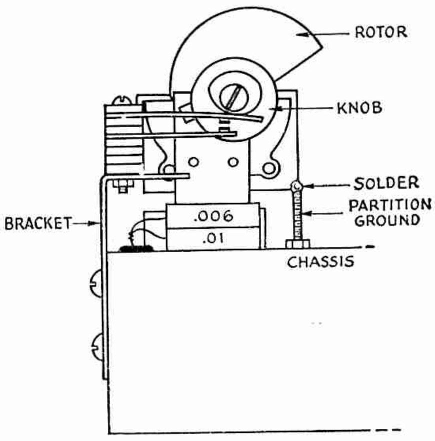


Fig. 5-36—The cam-operated switch, S_1, is made from the contacts and insulators taken from an open-circuit phone jack (Mallory 703) and mounted on an aluminum bracket. The cam, mounted on the shaft of C_1, is made by grinding one side of a small insulated knob (Johnson 116-214-1). Switch is open during minimum-capacitance half of capacitor range. Bracket is made from a 1¼ × 3½-inch strip of aluminum; the shelf is ¾-inch deep.

Construction

The receiver is built on an 8 × 12 × 3-inch aluminum chassis. A panel can be made from $\frac{1}{16}$-inch thick sheet aluminum or from a standard $8\frac{3}{4}$-inch rack panel. While the rack panel will be more substantial, it really isn't necessary, and the $\frac{1}{16}$-inch stock will be adequate. The panel is held to the chassis by the b.f.o. capacitor, C_8, the line/b.f.o. switch, S_2, the dial, and an extra pair of 6-32 screws.

It is worth while to mount the tuning capacitor, C_7, as accurately as possible with respect to the National ICN dial. For minimum backlash and maximum strength, C_7 is mounted on a three-sided aluminum housing that is securely fastened to the chassis on three sides by $\frac{3}{8}$-inch lips. A good flexible insulated coupling should be used between dial and capacitor shaft — a Millen 39006 is shown in the photograph.

The location of most of the major components can be determined by reference to the photographs. The inductors L_1L_2, L_3 and L_4L_5 are supported by suitable tie strips, as are the two 480-$\mu\mu$f. mica compression trimmers, C_2 and C_3, in the crystal filter circuit and the pair of 330-$\mu\mu$f. capacitors in the b.f.o. L_1L_2 should be wired so that the outside ends go to antenna and grid, and L_4L_5 should be wired with outside ends to plate and grid.

Details of the only unusual construction, the cam-operated switch S_1, are shown in Fig. 5-36. Note that the associated .006- and .01-μf. coupling capacitors are mounted above the chassis; a clearance hole with a rubber grommet is provided in the chassis for the common lead back to L_2 and L_3.

Since the rotor of C_1 must not make contact with the panel, a large clearance hole must be provided for the shaft bushing, and a pair of extruded fiber washers used to insulate the bushing from the panel. A brass screw or bayonet lug should be set into the chassis at the shield partition between the two stators of C_1, and the shield soldered to this chassis connection. The 3- to 30-$\mu\mu$f. compression trimmer across C_{1A} can be soldered between rotor and shield partition.

Many of the connections are made with shielded leads, to minimize hum and chances for feedback or feedthrough. The shielded leads are indicated in Fig. 5-35. The lead from the antenna jack is run in RG-58/U coaxial cable, as is the short lead from C_8 to a 330-$\mu\mu$f. capacitor. Heater leads to the tubes are made of shielded wire.

Alignment

The alignment procedure can be simplified if a short-wave receiver or a signal generator can be borrowed. Lacking these, a grid-dip meter can be used to provide a signal source and to check the resonances of the tuned circuits. If the 100-kc. oscillator can be checked on another receiver, it can be used to align the receiver. A broadcast receiver will tell if the 100-kc. oscillator is functioning — it should be possible to identify several of the oscillator's harmonics at 100-kc. intervals in the broadcast band, by the reduction in noise at those points.

The audio amplifier of the receiver can be checked by turning on the receiver and listening to the headphones as the audio gain control is advanced. When it is full clockwise a low-pitched hum should be just audible in the headphones. A further check can be made by bringing a finger near the arm of the audio gain control — the hum should increase.

If a means is available for checking the frequency of the b.f.o., it should be turned on at S_2 and set on or about 1700 kc. by means of the slug in L_8. Do this with C_8 set at half scale. If a broadcast receiver is the only measuring equipment you have, a 1700-kc. signal can be derived from it by tuning the receiver to 1245 kc., which puts its oscillator on 1700 kc. if the standard 455-kc. i.f. is used. A wire from around the receiver to the 2X4+1 should provide sufficient signal. Feeding a 1700-kc. signal into the detector by laying the source wire near the grid of the 6BA6 (i.f. gain arm at ground), it should be possible to peak L_7 for maximum signal and, as the signal frequency is changed slightly, a change in pitch of the whistle should be heard. With no incoming signal, a slight rushing noise should be heard in the head-phones when the b.f.o. is switched on by S_2. If this rushing noise is just barely discernible increase the capacitance at C_9 by adding a few more twists.

If the oscillator V_{1B} is operating, a voltmeter connected across the 4700-ohm 1-watt resistor in its plate lead should show an increase in voltage when the stator of C_5 or C_7 is shorted to ground momentarily with a screwdriver or other conductor. Connect the + lead of the voltmeter to the side of the resistor running to + 105 and the — connection to the .001-μf. side. If the oscillator doesn't work, it may be because the outside turns of L_4 and L_5 are not connected to plate and grid respectively. With the b.f.o. on and C_1 almost fully meshed, set the tuning capacitor C_7 at about 90 per cent full capacitance. Run C_5 to full capacitance and slowly reduce capacitance. At one point you should hear a loud signal, the second harmonic of the b.f.o. at 3400 kc. If the b.f.o. is reasonably close to frequency, turning on the calibration oscillator should give a weaker signal nearby (on the main tuning dial). Tune C_7 to a higher frequency (less capacitance) and you should hear another weaker signal, the 35th harmonic of the oscillator (3500 kc.). Peak C_1 for maximum signal and leave it. Run C_7 back to about 90 per cent full capacitance and then slowly reduce capacitance at C_5 until the 35th harmonic of the oscillator is again heard. If a 3500-kc. signal is available the adjustment can be made in a more straightforward manner.

Once the oscillator trimmer C_5 has been set to give the proper tuning range of the oscillator circuit (5.2 to 5.7 Mc.), the next problem is that of adjusting the crystal filter circuit. With a capacitance bridge, or a grid-dip meter and an inductance, are set the two capacitors C_2 and C_3 at the same capacitance (near maximum compression) before soldering them in the receiver. The actual value of capacitance isn't important. Lacking these instruments, tighten the capacitors to full compression and then loosen their screws by ¾ turn. Tune in a signal — it can be from the 100-kc. oscillator or any other steady source — and peak L_6 for maximum response. Tune off the signal until it disappears and set the pitch control, C_8, to a point where the background noise is reasonably high-pitched. This is easy to determine because at the lowest-pitched point there will be an increase in hum; make the lowest-pitched point the center of the knob scale by adjustment of L_8, and then set the pitch control to one end of its range. Tune back to the signal and "rock" the tuning, C_7, as you change the adjustment of L_6. Look for a condition that gives considerably more response on one side of zero beat than on the other. It is a good idea to buy several extra 1700-kc. crystals and try them in different combinations. Small changes in the setting of C_2 or C_3 will have an effect on the selectivity characteristic, but bear in mind that a change in C_2 or C_3 must be compensated for by a readjustment of L_6. With a little patience it should be possible to obtain a marked difference in the output strength on the two sides of zero beat. This will

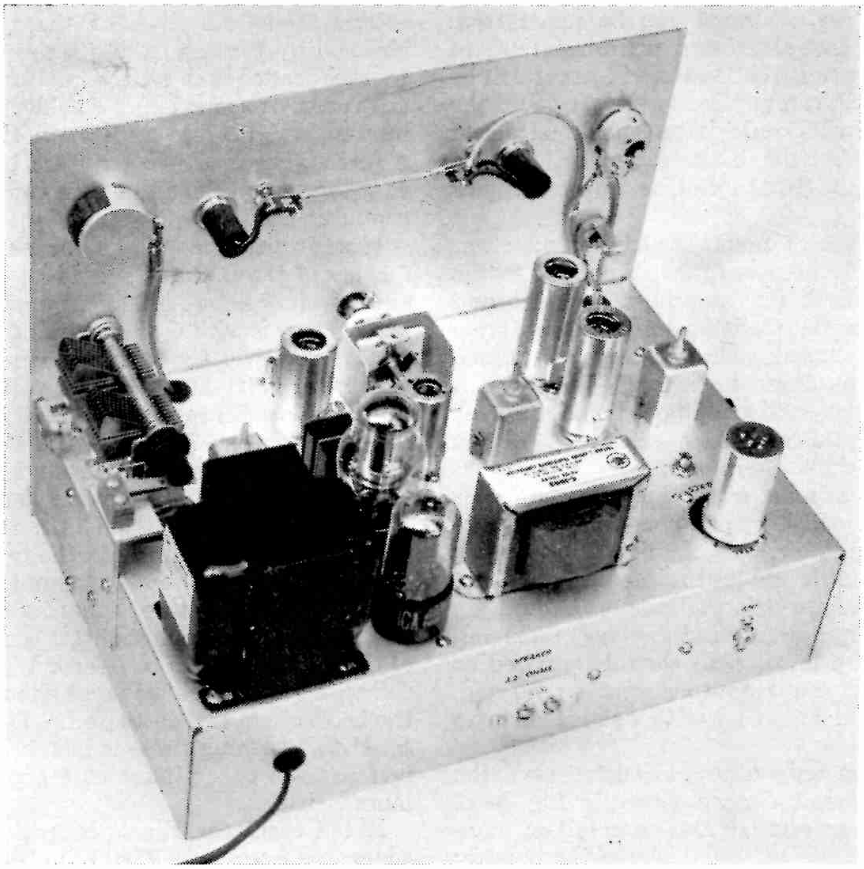


Fig. 5-37—Top view of the 2X4 + 1 receiver. The dual capacitor at the left tunes the receiver input; a homemake cam switch on its shaft changes the coupling between the two bands. The main tuning capacitor, rear center, is mounted on a three-sided aluminum bracket for maximum stability. The tube to the left of the bracket is the 6U8-A mixer-oscillator stage, and the 6BA6 i.f. stage is in front of the main tuning capacitor. The remaining tubes in shields are the 6CG7 detector/b.f.o. and the audio 6CG7 (near panel). Metal can plugged in socket above antenna jack houses 100-kc. calibrating crystal.

"flip over" to the other side if the pitch control is set at the other end of its range.

The remaining alignment job consists of bringing the input circuits into resonance on both bands. With a signal tuned in at 40 meters, "rock" C_1 back and forth to see if there are two (close-together) points where the signal peaks. If there are, adjust the 3-30-$\mu\mu$f. trimmer across L_2 until only one peak is found. Check on 80 meters in a similar fashion. If for any reason it is found that the two-peak condition can be eliminated on only one band at a time, it indicates an abnormal amount of antenna reactance, and a compromise adjustment will have to be made.

In operation, the receiver input control, C_1, should be set for maximum volume on the incoming signal or noise. The i.f. gain should be run at close to maximum on all but the loudest signals, and the audio gain control should be set for comfortable headphone or speaker volume. If an antenna changeover relay is used, it may be possible to monitor your own transmitter by detuning the input circuit to another band; this ability will depend upon the transmitter power and field in the vicinity of the receiver.

Frequency Standard

No trouble should be encountered with the 100-kc. oscillator if care is exercised in handling the transistor. When soldering its leads in place, hold the lead with a pair of pliers; the metal of the pliers will absorb heat and prevent injury to the transistor.

To tune the receiver to WWV, set C_7 to mid scale, set S_3 at the WWV position, peak C_1 on noise and slowly tune with C_6. On a busy day a wide variety of signals will be heard in this region; look for one with steady tone modulation and time ticks. If it can't be found within the range of C_6, set C_7 near one end of its range and try again. An alternate method is to disconnect the antenna, establish the position on the tuning dial (C_7) of several 100-kc. harmonics, connect the antenna and investigate each one of these frequencies. Depending upon one's geographical location, there will be times when WWV cannot be heard on 5 Mc., so don't be discouraged by failure on the first try. Once WWV has been located with good strength, the 50th harmonic of the 100-kc. crystal can be brought to zero beat with WWV by adjustment of C_4.

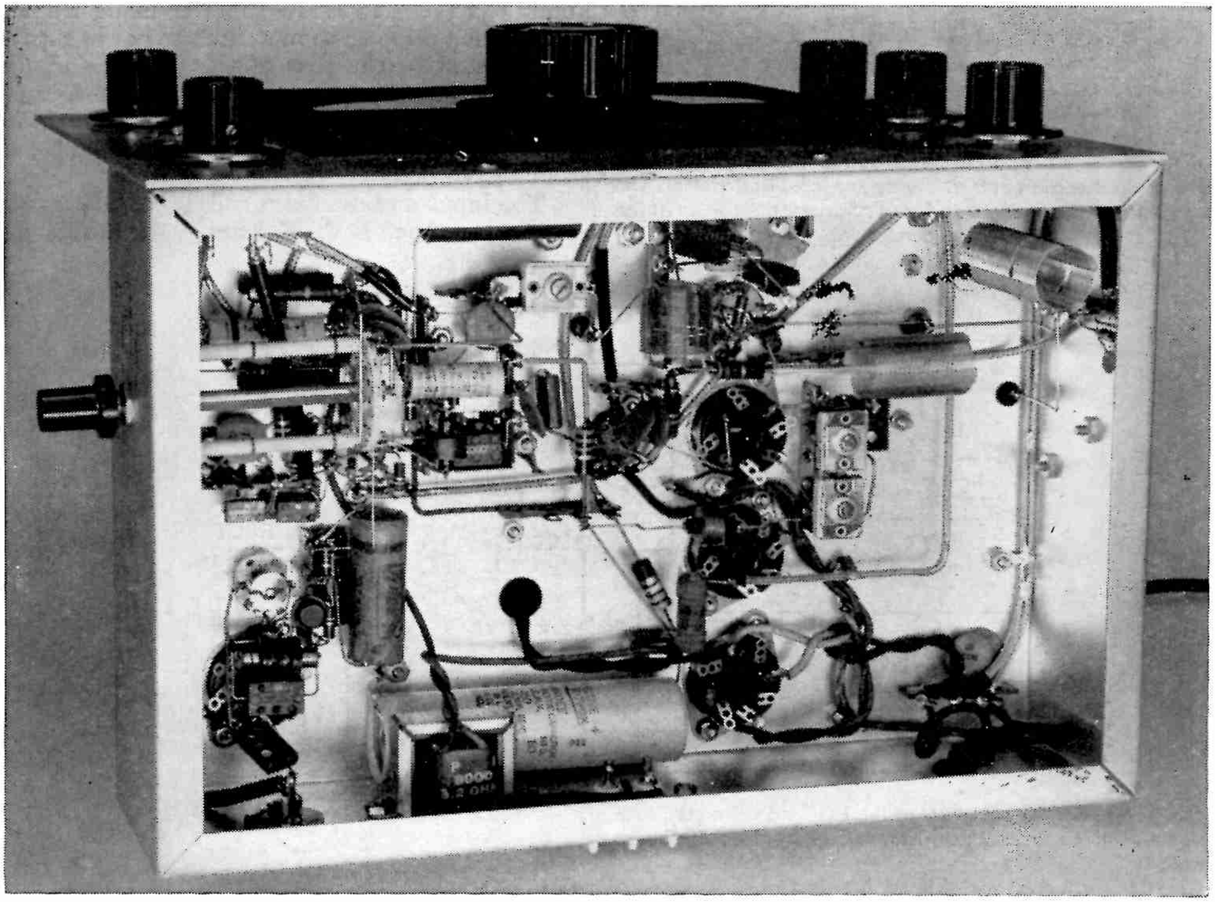


Fig. 5-38—The input inductors L_1L_2 are supported by a terminal strip on the side of the chassis (upper right), and L_3 is supported nearby by a terminal strip mounted on the chassis. The coils are at right angles to minimize inductive coupling. The oscillator inductors, L_4L_5, are also supported by a terminal strip (top center). A mica compression trimmer to the left of the oscillator inductors is used to center WWV on the tuning dial; the pair of compression trimmers below L_3 are in the crystal filter circuit.

A Selective Converter for 80 and 40 Meters

Many inexpensive "communications" receivers are lacking in selectivity and bandspread. The 80- and 40-meter performance of such a receiver can be improved considerably by using ahead of it the converter shown in Figs. 5-39 and 5-41. This converter is not intended to be used ahead of a broadcast receiver except for phone reception, because the BC set has no b.f.o. or manual gain control, and both of these features are necessary for good c.w. reception. The converter can be built for less than $20, and that cost can be cut appreciably if the power can be "borrowed" from another source.

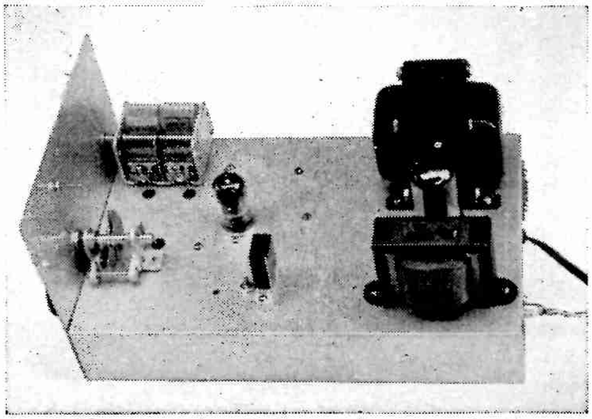


Fig. 5-39—Used ahead of a small receiver that tunes to 1700 kc., this converter will add tuning ease and selectivity on the 80- and 40-meter bands. The input capacitor is the dual section unit at the upper left-hand corner. The crystal and the tuning slug for L_6 are near the center at the foreground edge.

The converter uses the tuning principle employed in the two-band superheterodynes described earlier in this section. A double-tuned input circuit with large capacitors covers both 80 and 40 meters without switching, and the oscillator tunes from 5.2 to 5.7 Mc. Consequently with an i.f. of 1700 kc. the tuning range of the converter is 3.5 to 4.0 Mc. and 6.9 to 7.4 Mc. Which band is being heard will depend upon the setting of the input circuit tuning (C_1 in Fig. 5-40). The converter output is amplified in the receiver, which must of course be set to 1700 kc. To add selectivity, a 1700-kc. quartz crystal is used in series with the output connection. A small power supply is shown with the converter, and some expense can be eliminated if 300 volts d.c. at 15 ma. and 6.3 volts a.c. at 0.45 ampere is available from an existing supply.

Construction

The unit is built on a 7 × 11 × 2-inch aluminum chassis. The front panel is made from a 6 × 7-inch piece of aluminum. The power supply is mounted to the rear of the chassis and the converter components are in the center and front. The layout shown in the bottom view should be followed, at least for the placement of L_1, L_2, L_3 and L_4.

The input and oscillator coils are made from a single length of B & W Miniductor stock, No.

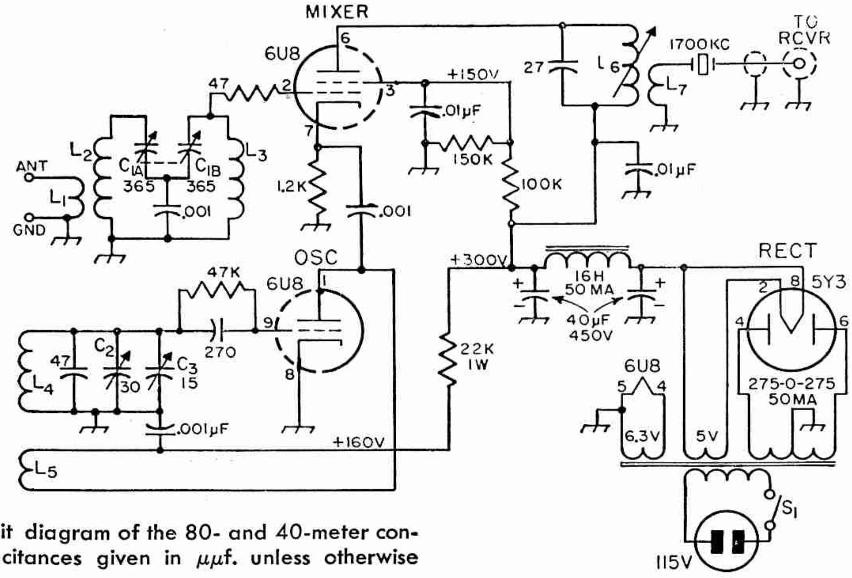


Fig. 5-40—Circuit diagram of the 80- and 40-meter converter. All capacitances given in μμf. unless otherwise noted.

C_1—365-μμf. dual variable, t.r.f. type.
C_2—3–30-μμf. trimmer.
C_3—15-μμf. variable (Bud 1850, Cardwell ZR-15AS, Millen 20015).
L_1, L_2, L_3, L_4, L_5—B & W No. 3016 Miniductor, 1-inch diameter, 32 turns per inch, No. 22 wire, cut as below.
L_1—8 turns separated from L_2 by one turn (see text).
L_2, L_3—19 turns.
L_4—21 turns separated from L_5 by one turn.
L_5—8 turns.
L_6—105–200-μh. slug-tuned coil (North Hills Electric 120H).
L_7—See text.
Crystal—1700 kc. (E. B. Lewis Co. Type EL-3).

3016. Count off 31 turns of the coil stock and bend the 32nd turn in toward the axis of the coil. Cut the wire at this point and then unwind the 32nd turn from the support bars. Using a hacksaw blade, carefully cut the polystyrene support bars and separate the 31-turn coil from the original stock. Next, count off 9 turns from the 31-turn coil and cut the wire at the 9th turn. At the cut unwind a half turn from each coil, and also unwind a half turn at the outside ends. This will leave two coils on the same support bars, with half-turn leads at their ends. One coil has 21 turns and the other has 8 turns, and they are separated by the space of one turn. These coils are L_4 and L_5.

The input coils L_1 and L_2 are made up in the same manner. Standard bakelite tie points are used to mount the coils. Two 4-terminal tie points are needed for L_1L_2 and L_4L_5, and a one-terminal unit is required for L_3. The plate load inductance L_6 is a 105–200 μh. variable-inductance coil (North Hills 120H). The coupling coil L_7 is 45 turns of No. 32 enam. scramble-wound adjacent to L_6. If the constructor should have difficulty in obtaining No. 32 wire, any size small enough to allow 45 turns on the coil form can be substituted.

The input capacitor, C_1, is a 2-gang t.r.f. variable, 365 $\mu\mu$f. per section. As both the stators and rotor must be insulated from the chassis, extruded fiber washers should be used with the screws that hold the unit to the chassis. The panel shaft hole should be made large enough to clear the rotor shaft.

A National type O dial assembly is used to tune C_3. One word of advice when drilling the holes for the dial assembly: the template furnished with the unit is in error on the 2-inch dimension (it is slightly short) so use a ruler to measure the hole spacing.

In wiring the unit, it is important that the output lead from the crystal socket be run in shielded wire. A phono jack is mounted on the back of the chassis, and a piece of shielded lead connects from the jack to the crystal socket terminal. The leads from the stators of C_1 and C_3 are insulated from the chassis by means of rubber grommets.

Testing and Adjustment

A length of shielded wire is used to connect the converter to the receiver: the inner conductor of the wire is connected to one antenna terminal; the shield is connected to the other terminal and grounded to the receiver chassis. The use of shielded wire helps to prevent pickup of unwanted 1700-kc. signals. Turn on the converter and receiver and allow them to warm up. Tune the receiver to the 5.2-Mc. region and listen for the oscillator of the converter. The b.f.o. in the receiver should be turned on. Tune around until the oscillator is heard. Once you spot it, tune C_3 to maximum capacitance and the receiver to as close to 5.2 Mc. as you can. Adjust the oscillator trimmer capacitor, C_2, until you hear the oscillator signal. Put your receiving antenna on the converter, set the receiver to 1700 kc., and tune the input capacitor, C_1, to near maximum capacitance. At one point you'll hear the background noise come up. This is the 80-meter tuning. The point near minimum capacitance — where the noise is loudest — is the 40-meter tuning.

With the input tuning set to 80 meters, turn on your transmitter and tune in the signal. By spotting your crystal-controlled frequency you'll have one sure calibration point for the dial. By listening in the evening when the band is crowded you should be able to find the band edges.

You'll find by experimenting that there is one point at or near 1700 kc. on your receiver where the background noise is the loudest. Set the receiver to this point and adjust the slug on L_6 for maximum noise or signal. When you have the receiver tuned *exactly* to the frequency of the crystal in the converter, you'll find that you have quite a bit of selectivity. Tune in a c.w. signal and tune slowly through zero beat. You should notice that on one side of zero beat the signal is strong, and on the other side you won't hear the signal or it will be very weak (if it isn't, off-set the b.f.o. a bit). This is single-signal c.w. reception.

When listening to phone signals, it may be found that the use of the quartz crystal destroys some of the naturalness of the voice signal. If this is the case, the crystal should be unplugged and replaced by a 10- or 20-$\mu\mu$f. capacitor.

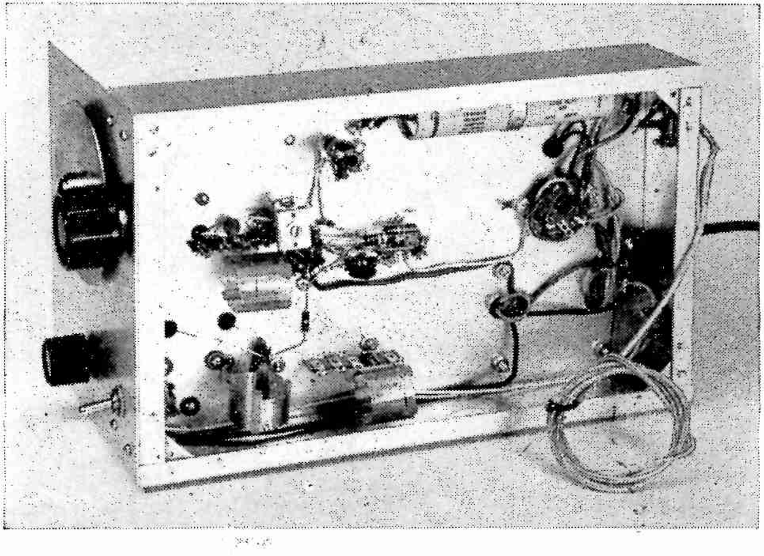


Fig. 5-41—Bottom view of the converter showing placement of parts. The coil at the lower left is L_3, and the input coil, L_1L_2, is just to the right of L_3. The oscillator coil L_4L_5, is at the left near the center. The output coil, L_6, is near the top center.

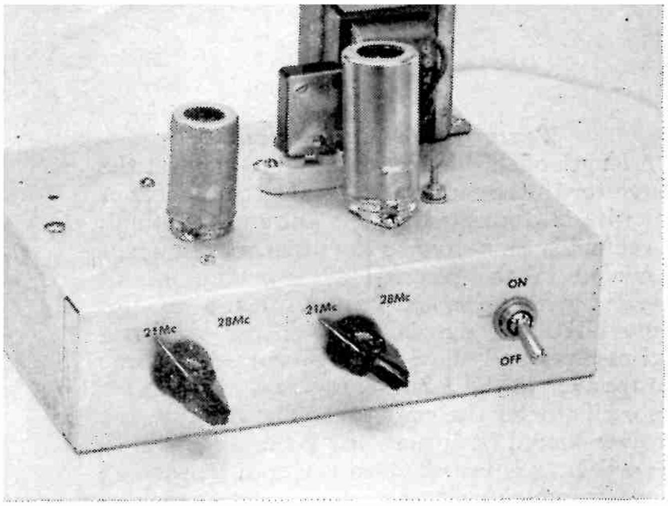


Fig. 5-42—This view of the "bonus" converter shows all of the components projecting above the chassis. At the left on the front is the r.f. control and next to it is the mixer tuning. At the far right is the a.c. switch. The tube at the left is the r.f. amplifier, and the crystal is between the transformer and the mixer tube. Screw adjustment to the right of the mixer tube sets the slug of L_5.

The "Bonus" 21-Mc. Converter

The cure for most of the high-frequency ills of many receivers is the installation of a good crystal-controlled converter between the antenna and the receiver. The converter shown in Figs. 5-42 and 5-43, while intended primarily for 21-Mc. operation, gives a bonus of 28-Mc. reception without any additional parts or switching. This is accomplished by using signal circuits that tune more than the 21- to 30-Mc. range and using a crystal-controlled oscillator at 25 Mc. Using the converter ahead of a receiver, the 15-meter band, 21.0 to 21.45 Mc., will be found from 4.0 to 3.55 on the receiver. The receiver tunes "backwards." The 10-meter band tunes 3.0 to 4.7 Mc. on the receiver.

Referring to Fig. 5-44, the converter consists of three stages, but it uses only two tubes. An r.f. stage amplifies the incoming signals, and an oscillator provides a steady signal that, in a mixer stage, heterodynes the incoming signal to the *difference* frequency mentioned above. If the input and output circuits of the r.f. stage aren't tuned to 21 Mc. the 21-Mc. signals can't be amplified to the full capability of the stage. However, the 21-Mc. tuned circuits aren't too sharp, so a single-setting will usually suffice for most of the 21-Mc. band, and all of the tuning will normally be done at the receiver alone. The 47 000-ohm resistor across C_2 was used to make the associated circuit a bit broader.

The selenium-rectifier power supply is quite adequate for the job and makes the converter a self-sufficient unit, although the power may be "borrowed" from the receiver if it is felt that the selenium supply is an unnecessary expense.

In the crystal-controlled oscillator portion, a capacitive divider (C_3 and C_4) provides a tap on the tank circuit so that the oscillator is loaded very lightly. If you didn't tap down on the tuned circuit the overtone crystal, Y_1, might show lower-frequency energy as well, or it might not oscillate at all.

The size of the chassis shown in Figs. 5-42 and 5-43 is 2 × 5 × 7 inches. However, any chassis large enough to accommodate the parts can be used. Most of the construction is simple but there are a few places where certain precautions should be taken, and these will be treated in detail.

Study the photographs, particularly the bottom view, to see how the coils and tube socket are mounted. Notice the shield that cuts across the 6AK5 socket. The purpose of the shield is to minimize the coupling between the grid and plate circuits of the r.f. stage, to avoid oscillation. A scrap of roofing copper was cut to 3½ by 2 inches for the shield. Brass, or any other metal that can be soldered, could be substituted. The shield and socket should be mounted so that the shield bisects the socket between Pins 4 and 5. There is a ¼-inch lip on the shield which is used to mount it to the chassis top. The metal tube in the center of the tube socket should be soldered to the shield; the shield is held to the chassis by two 6-32 screws. Soldering lugs should be mounted under

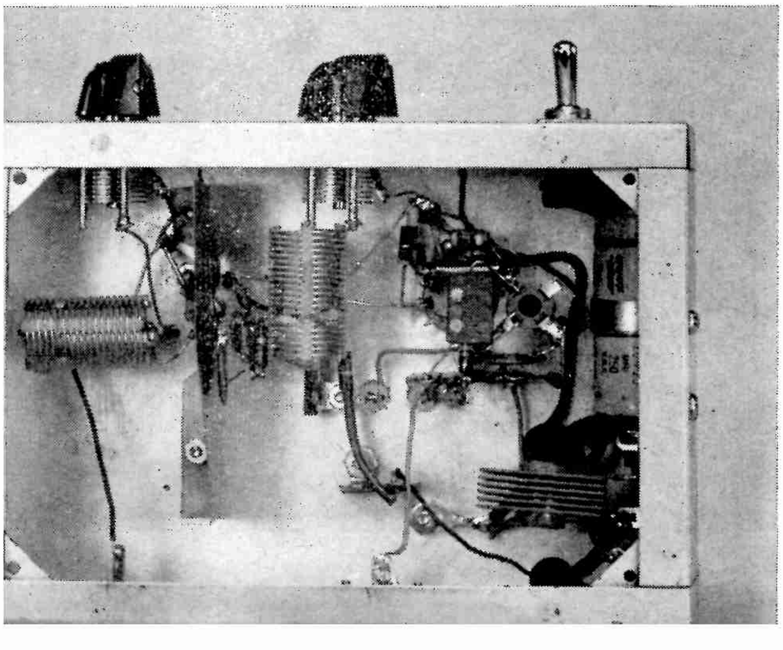


Fig. 5-43—All of the components of the power supply are grouped at the right. The tubular capacitor, C_5, mounts against the chassis wall. At the opposite side of the chassis, the metal strip shields the input circuit of the r.f. stage. The coils to the right of the shield are L_3, and L_4.

21 Mc. Converter

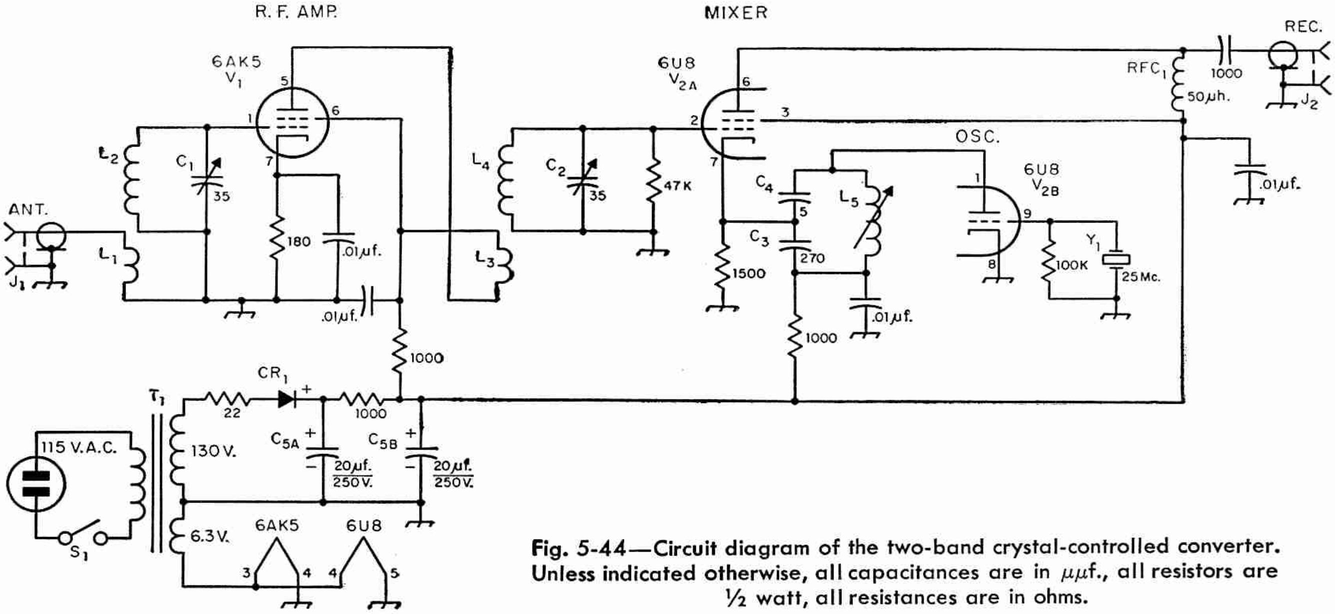


Fig. 5-44—Circuit diagram of the two-band crystal-controlled converter. Unless indicated otherwise, all capacitances are in µµf., all resistors are ½ watt, all resistances are in ohms.

C_1, C_2—35-µµf. midget variable (Hammarlund MAPC-35-B).
C_3—270-µµf. silver mica or NPO ceramic.
C_4—5-µµf. silver mica or NPO ceramic.
C_5—Dual electrolytic, 20-20 µf. at 250 volts.
CR_1—100-ma. 150-volt selenium rectifier (International Rectifier RS-100-E or equiv.).
J_1, J_2—Phono jack, RCA style.
L_1, L_2, L_3, L_4—Made of No. 20 bare, ⅝-inch diam., 16 t.p.i. stock. See text. (B & W Miniductor No. 3007).
L_5—2- to 3-µh. slug-tuned inductor (North Hills 120-A).
RFC_1—50-µh. r.f. choke (National R-33, Millen 34300-50).
S_1—S.p.s.t. toggle.
T_1—125 volts at 50 ma., 6.3 volts at 2 amperes (Stancor PA8422) or 135 volts at 50 ma., 6.3 volts at 1.5 amperes (Triad R-30-X).
Y_1—25.00-Mc. crystal (International Crystal Co., type FA-9).

the nuts that hold the 6AK5 socket, and all the chassis ground connections of the 6AK5 grid and plate circuit should be made to these lugs.

The coils are made from B & W 3007 Miniductor stock. To make the coils, first cut off a coil of 21 turns from the stock. Next, unwind one turn from each end of the 21-turn coil. Now count off 5½ turns from one end and cut the wire at this point. If you bend the 4th and 6th turns in toward the center of the coil you should be able to reach the 5th turn with your wire cutters. Unwind the half turn from each side leaving two coils on the same support bars, one 5 turns and the other 13 turns. Two of these dual coils are needed, one for the r.f. stage and the other for the mixer. They can be mounted on a standard terminal tie point or supported by their own leads. Tie points provide a more rigid support.

The power supply is a simple half-wave rectifier, using a transformer, selenium rectifier, and an *RC* filter circuit. Incidentally, when connecting the rectifier, the + side is connected to the *output* side of the supply. Again, a standard terminal tie point is used for most of the connections of the supply.

The preliminary checks are simple and should present no problems to the builder. First, turn on S_1 and see if the tubes light up. If they don't, turn off the switch and carefully check the wiring. Once the tubes light, allow a minute or two for the unit to warm up. The first thing to check is the crystal-controlled oscillator. If your receiver tunes to 25 Mc., listen in that region for the oscillator signal, which should come in loud and clear. If it doesn't, adjust the slug of L_5 until the oscillator starts. Should you find that it doesn't oscillate you'll need to make some voltage checks to make sure there is plate voltage on the oscillator. The voltage should be approximately 110, give or take 10 volts. If no voltage is indicated, check the wiring for errors.

Connect the converter to your receiver, using a piece of coax as the connecting line. Coax is used for the lead between the two units to minimize any pickup of unwanted signals near or in the 80-meter band. Set your receiver to tune the right range, 4000 to 3550 kc., and turn both units on.

Adjust C_1 and C_2 for maximum background noise. You'll find two values of capacitance (four points) on each capacitor that will give an increase in noise, one near minimum capacitance (plates unmeshed) and the other with more capacitance. The setting at the greater capacitance point is 21 Mc. while the lesser is 28 Mc. Adjust the converter for maximum noise at 21 Mc. and tune your receiver across the band. If the band is open — and don't forget that sometimes it's as dead as the famous doornail — you should hear signals. Tune in one and peak it up by tuning C_1 and C_2 of the converter. Each control should give a definite peak. Pretty nice to know that your receiving front end is lined up, isn't it? And it is, you know; you align it when you peak the two controls. Your receiver is now working as a tunable i.f. and the only adjustment required is to peak the antenna trimmer (if you have one) for maximum signal.

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 ampere. 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 μf.

The wiring diagram of the Selectoject is shown in Fig. 5-45. Resistors R_2 and R_3, and R_4 and R_5, can be within 10 per cent 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. 5-45. 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-μf. electrolytic capacitor 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.

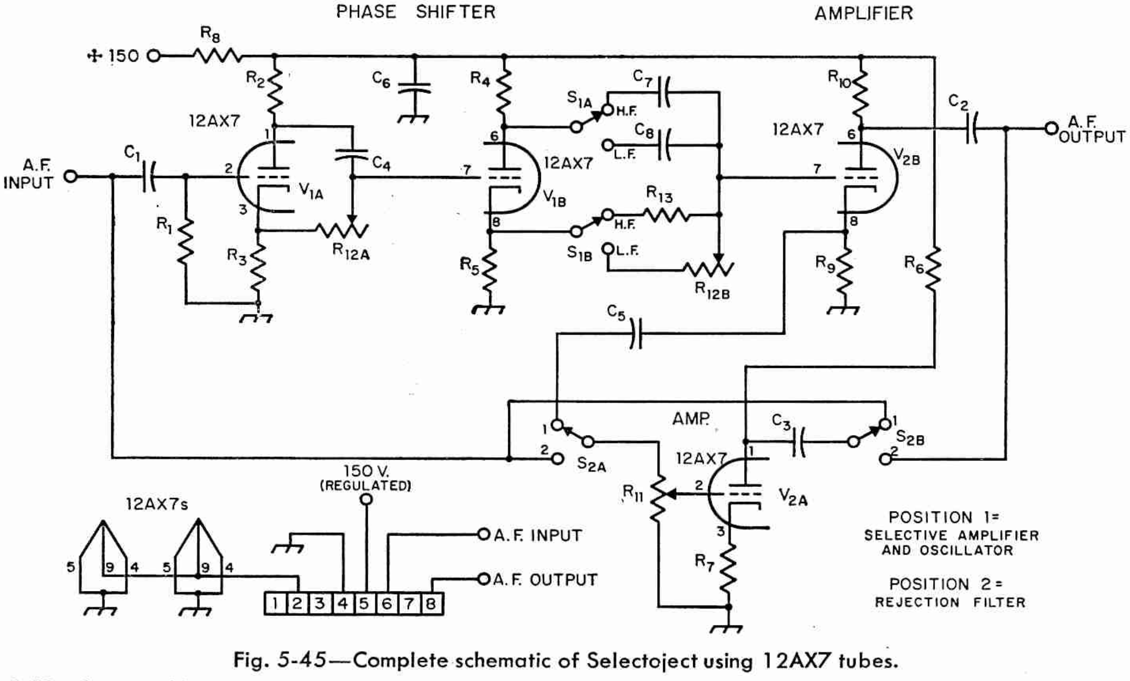


Fig. 5-45—Complete schematic of Selectoject using 12AX7 tubes.

C_1—0.01-μf. mica, 400 volts.
C_2, C_3—0.1-μf. paper, 200 volts.
C_4, C_8—0.002-μf. paper, 400 volts.
C_5—0.05-μf. paper, 400 volts.
C_6—16-μf. 150-volt electrolytic.
C_7—0.0002-μf. mica.
R_1—1 megohm, ½ watt.
R_2, R_3—1000 ohms, 1 watt, matched as closely as possible (see text).
R_4, R_5—2000 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}—0.5-megohm ½-watt potentiometer (selectivity).
R_{12}—Ganged 5-megohm potentiometers (tuning control) (IRC PQ11-141 with IRC M11-141.)
R_{13}—0.12 megohm, ½ watt.
S_1, S_2—D.p.d.t. toggle (can be ganged).

Antenna Coupler for Receiving

In many instances reception can be improved by the addition of an antenna coupler between the antenna feedline and the receiver, and in all cases the r.f. image rejection will be increased. The unit shown on this page consists of one series-tuned circuit and one parallel-tuned circuit; usually its best performance is obtained with the parallel-tuned circuit connected to the receiver input, as indicated in Fig. 5-46. However, the coupler should also be tried with the connections reversed, to see which gives the better results. The desired connection is the one that gives the sharper peak or louder signals when the circuits are resonated.

The coupler is built on one section of a 5 × 4 × 3-inch Minibox (Bud CU-2105A). Tuning capacitors C_1 and C_2 are mounted directly on the Minibox face, since there is no need to insulate the rotors. The arrangement of the components can be seen in Fig. 5-47.

The coils L_1 and L_2 are made from a single length of B & W 3011 Miniductor. The wire is snipped at the center of the coil and unwound in both directions until there are three empty spaces on three support bars and two empty spaces on the bar from which the snipped ends project. These inner ends run to the connectors J_1 and J_2. (Fig. 5-46). Unwind turns at the ends of the coils until each coil has a total of 22 turns. When soldering the leads to the 3rd, 6th, 8th and 12th turns from the inside ends of the coils, protect the adjacent turns from solder and flux by placing strips of aluminum cooking foil between the turns. An iron with a sharp point will be required for the soldering.

The "panel" side of the box can be finished off with decals indicating the knob functions and switch positions.

The antenna coupler should be mounted within a few feet of the receiver, to minimize the length of RG-59/U between coupler and receiver. In crowded quarters, the use of M-359A right-angle adapters (Amphenol 83–58) and J_1 and J_2 will make it possible to bring out the cables in better lines.

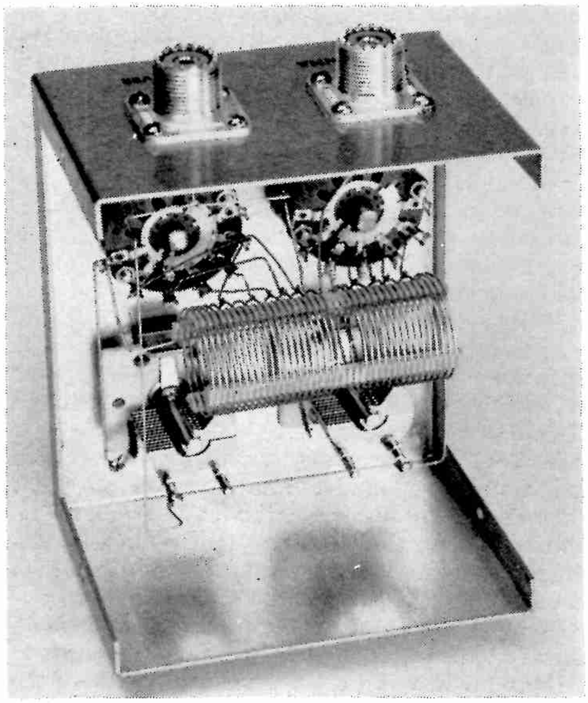


Fig. 5-47—Receiver antenna coupler, with cover removed from case. Unit tunes 6 to 30 Mc. The coil is supported by the leads to the capacitors and switches.

Normally the coupler will be adjusted for optimum coupling or maximum image rejection, but by detuning the coupler it can be used as an auxiliary gain control to reduce the overloading effects of strong local signals. The coupler circuits do not resonate below 6 Mc., but a coupler of this type is seldom if ever used in the 80-meter band; its major usefulness will be found at the higher frequencies.

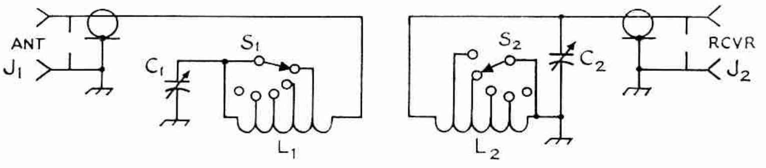


Fig. 5-46—Circuit diagram of the receiver antenna coupler.

C_1, C_2—100-$\mu\mu$f. midget variable (Hammarlund HF-100).

J_1, J_2—Coaxial cable connector, SO-239.

L_1, L_2—22 turns No. 20, ¾-inch diameter, 16 t.p.i. Tapped 3, 6, 8 and 12 turns from inside end. See text on spacing and tapping.

S_1, S_2—Single-pole 11-position switch (5 used) rotary switch (Centralab PA-1000).

A Regenerative Preselector for 7 to 30 Mc.

The performance of many receivers begins to drop off at 14 Mc. and higher. The signal-to-*noise ratio* is reduced, and unless double conversion is used in the receiver there is likely to be increased trouble with r.f. images at the higher frequencies. The preselector shown in Figs. 5-48 and 5-49 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. The poorer the receiver is at the higher frequencies, the more it will benefit by the addition of the preselector.

A truly good receiver at 28 Mc. will show little or no improvement when the preselector is added, but a mediocre receiver or one without an r.f. stage will be improved greatly through the use of the preselector.

A 6CG7 dual triode is used in the preselector, one triode as a band switched regenerative r.f. stage and the other as a cathode follower. A conventional neutralizing circuit is used in the amplifier; by upsetting this circuit enough the stage can be made to oscillate. Smooth control of regeneration up to this point is obtained by varying one of the capacitances in the neutralizing circuit.

If and when it becomes necessary to reduce gain (to avoid overloading the receiver), the regeneration control can be retarded. One position of the bandswitch permits straight-through operation, so the preselector unit can be left connected to the receiver even during low-frequency reception.

The preselector is built on a 5 × 10 × 3-inch chassis (Bud AC-404). A 5× 6½-inch aluminum panel is held to the chassis by the extension-shaft bushing for the regeneration-control capacitor, C_3, and the bushing for the rotary switch. The coils, L_1 and L_2, are supported on a small staging of 1¼ × 3-inch clear plastic. (It can be made from the lid of the box that the Sprague 5GA-S1 .01-μf. disk ceramic capacitors come in.) All coils can be made from a single length of B&W 3011 Miniductor. They are cemented to the plastic staging with Duco cement.

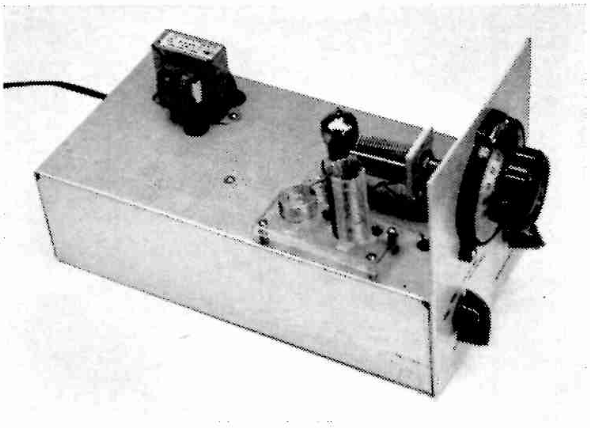


Fig. 5-48—The regenerative preselector covers the range 7 to 30 Mc.; it can be used ahead of any receiver to improve gain, image rejection and, in many cases, sensitivity. A dual triode 6CG7 is used as r.f. amplifier and cathode follower.

The rotor of C_1 can be insulated from the chassis by mounting the capacitor bracket on insulating bushings (National XS-6 or Millen 37201); its shaft is extended through the use of an insulated extender shaft (Allied Radio No. 60 H 355). The bandswitch S_1 is made from the specified sections (see Fig. 5-50).

The first section is spaced ¾ inch from the indexing head, there is 1-inch separation be-

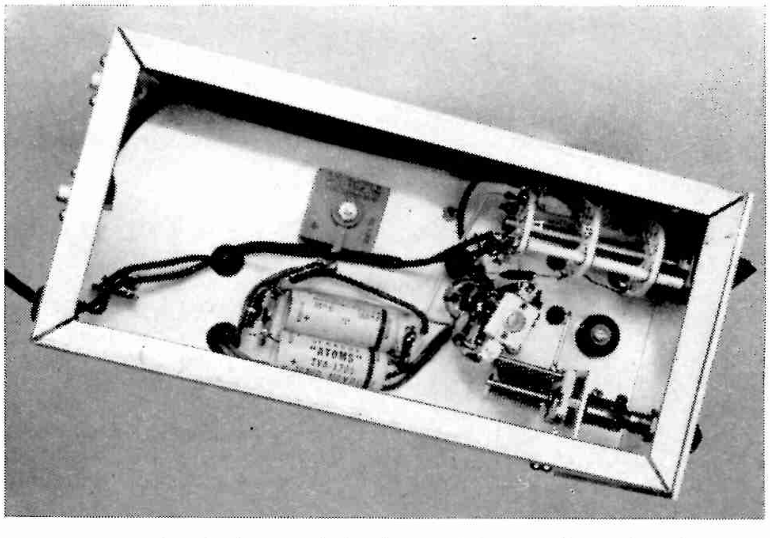


Fig. 5-49—The r.f. components are bunched around the 9-pin miniature tube socket. Power supply components are supported by screws and tie points.

Regenerative Preselector

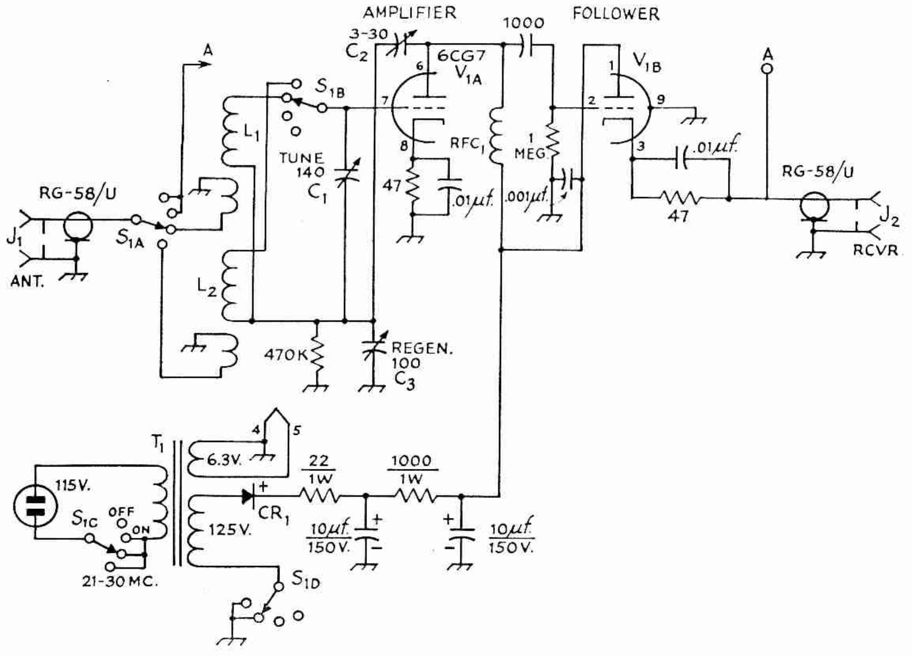


Fig. 5-50—Circuit diagram of the regenerative preselector. Unless otherwise specified, resistors are ½ watt, capacitors are in μμf., capacitors marked in polarity are electrolytic.

C_1—140-μμf. midget variable (Hammarlund HF-140).

C_2—3- to 30-μμf. mica compression trimmer.

C_3—100-μμf. midget variable (Hammarlund MAPC-100-B).

CR_1—50-ma. selenium rectifier (International Rectifier RSO50).

J_1, J_2—Phono jack.

L_1—19 turns, 7-turn primary.

L_2—5 turns, 2-turn primary. Coils are ¾-inch diameter, 16 t.p.i., No. 20 wire (B & W 3011 Miniductor). One-turn spacing between coils and primaries.

S_1—Three-wafer switch. S_{1A} and S_{1B} are 1-pole 12-position (4 used) miniature ceramic switch sections (Centralab PA-1); S_{1C} and S_{1D} are 2-pole 6-position (4 used) miniature switch (Centralab PA-3). Sections mounted on Centralab PA-301 index assembly.

T_1—125 v. at 15 ma., 6.3 v. at 0.6 amp. (Stancor PS-8415).

RFC_1—100-μh. r.f. choke (National R-33).

tween this and the next section (S_{1B}), and the next section (S_{1C}, S_{1D}) is spaced 2½ inches from S_{1B}.

The regeneration control, C_3, is mounted on a small aluminum bracket. Its shaft does not have to be insulated from the chassis, so either an insulated or a solid shaft connector can be used. The small neutralizing capacitor, C_2, is supported by soldering one lead of it to a stator bar of C_3 and running a wire from the other lead to pin 6 of the tube socket. The rotor and stator connections from C_1 are brought through the chassis deck through small rubber grommets.

Power supply components, resistors and capacitors are supported by suitable lugs and tie points. Phono jacks are used for the input and output connectors.

Adjustment

Assuming that the wiring is correct and that the coils have been constructed properly and cover the required ranges, the only preliminary adjustment is the proper setting of C_2. Connect an antenna to the input jack and connect the receiver to the output jack through a suitable length of RG-58/U. Turn on the receiver b.f.o. and tune to 28 Mc. with S_1 in the ON position. Now turn S_1 to the 21- to 30-Mc. range. Swing the TUNING capacitor, C_1, and listen for a loud rough signal which indicates that the preselector is oscillating. If nothing is heard, advance the regeneration control toward the minimum capacitance end and repeat. If no oscillation is heard, it may be necessary to change the setting of C_2. Once the oscillating condition has been found, set the regeneration control at minimum capacitance and slowly adjust C_2 until the preselector oscillates only when the regeneration control is set at minimum capacitance. You can now swing the receiver to 21 Mc. and peak the preselector tuning capacitor. It will be found that the regeneration capacitance will have to be increased to avoid oscillation.

Check the performance on the lower range by tuning in signals at 14 and 7 Mc. and peaking the preselector. It should be possible to set the regeneration control in these two ranges to give both an oscillating and a non-oscillating condition of the preselector.

A little experience will be required before you can get the best performance out of the preselector. Learn to set the regeneration control so that the preselector is selective, but not so selective that it must be retuned every 10 kc. or so. Changing antenna loads will modify the correct regeneration control setting.

A Clipper/Filter for C.W. or Phone

The clipper/filter shown in Fig. 5-51 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 and it will keep the strength of c.w. signals at a constant level, and while the filter will add selectivity to your receiver for c.w. reception, the unit 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.

There are times when only the selective audio circuits will be wanted, while on other occasions only clipping will be needed. Since it is a simple matter to provide a switching arrangement so that either function, or both, can be used at will, this has been done in the unit described here.

The frequency response of the selective circuits reaches a peak at about 700 cycles and has a null at about 2000 cycles. The peak frequency is determined by the combined values of L_1, C_1, and C_2 (or L_2, C_3 and C_4), while the notch frequency is that of the parallel-resonant circuit L_1C_1 (or L_2C_3). If different peak and null frequencies are desired the values of C_1 and C_2 (and C_3 and C_4) can be changed; for raising the notch frequency the capacitance of C_1 and C_3 should be made smaller; to raise the peak frequency reduce the capacitance at C_2 and C_4.

The rotary switch S_2 (Fig. 5-51) is used to provide different combinations of the clipper and filter. To simplify the wiring diagram the switching circuit is shown separately in the diagram.

The filter-clipper can be built on an aluminum chassis, but a steel cabinet should be used to house the unit. Steel is preferable to aluminum because L_1 and L_2 are sensitive to stray magnetic fields (which would show up as hum at the output) and the steel cabinet aids in shielding. One layout precaution should be observed: Place the filter inductors L_1 and L_2 as far as possible from the power transformer, and mount them with their cores at right angles to the core of the transformer. This will minimize hum pickup by the inductors.

Before mounting L_1 and L_2, it will be necessary to remove the mounting frames and insulate the "I" laminations, as shown in Fig. 5-52. The frame is removed easily by prying out its two legs and then lifting it from the core. The "I" laminations are in the form of a bar lying across the top of the "E" core.

By mounting the chokes with nonmetallic straps the Q will remain high. If aluminum or other nonmagnetic materials are used the Q will

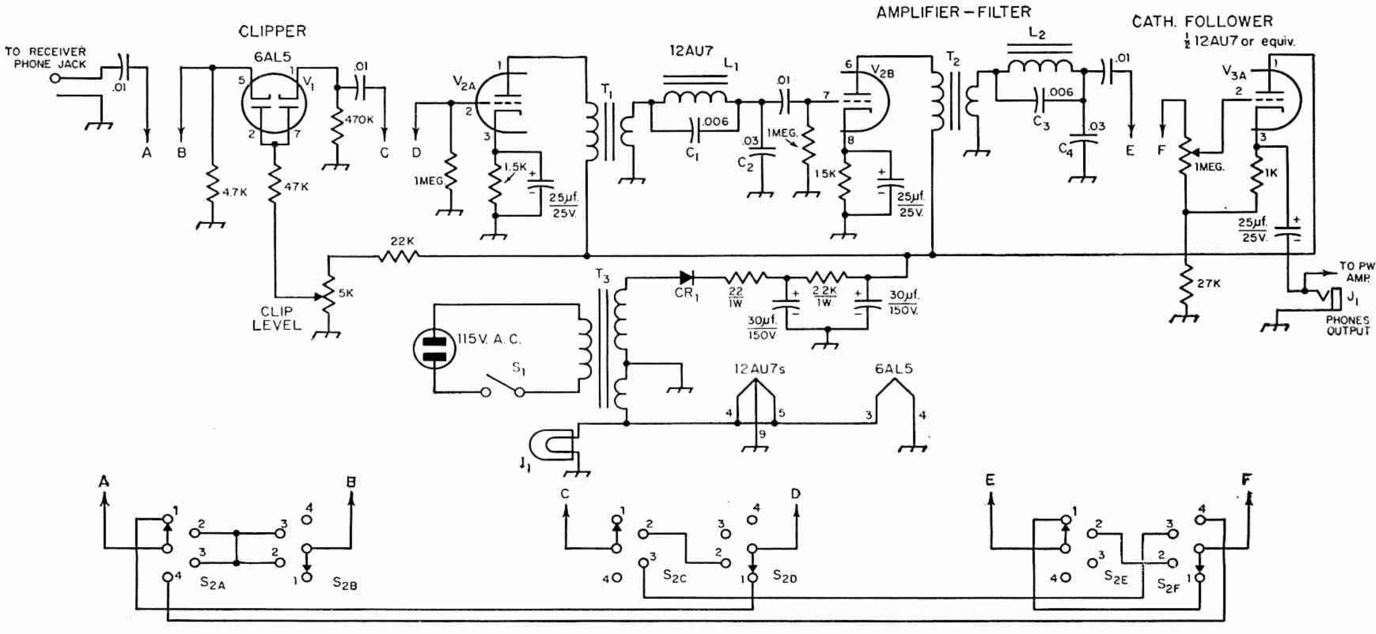


Fig. 5-51—Circuit of the two-stage clipper-filter. All capacitances are in µf. All 0.01 µf. capacitors may be ceramic; capacitors marked with polarity are electrolytic. Others should be tubular plastic or mica. Resistors are ½ watt unless otherwise specified. Switch functions are as follows: Position 1, dual filter alone; Position 2, clipper and dual filter; Position 3, clipper alone; Position 4, straight through with cathode-follower output.

CR_1—50-ma. selenium rectifier.
I_1—6.3-volt pilot lamp.
J_1—Open-circuit phone jack.
L_1, L_2—5-h. 65-ma. filter choke; frame removed and choke remounted as described in the text.
S_1—S.p.s.t. toggle switch
S_2—3-section 6-pole 4-position rotary switch, shorting type preferable. (Centralab PA-1020).
T_1, T_2—Output transformer: 7000—10,000-ohm primary to 3.2-ohm voice coil (Thordarson24S52).
T_3—Power transformer: 125 volts, 50 ma.; 6.3 volts, 2 amps. (Stancor PA-8421).

A Clipper/Filter

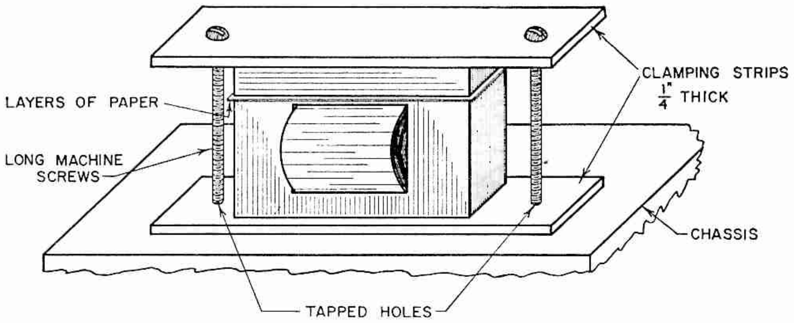


Fig. 5-52—Sketch showing the method of clamping and tuning the filter inductors. Clamping strips must be of bakelite, phenol, plastic or other suitable *insulating* material. Metal should not be used.

be adversely affected and the selectivity of the filter will suffer.

The switch wiring shown at the bottom of the schematic diagram can be done before mounting S_2 in place. After the switch is mounted the wiring between it and the other components can be completed.

Apply power by closing S_1, insert the plug in the receiver phone jack and turn switch S_2 to the "out" or straight-through position. Tune the receiver until a c.w. signal is found and adjust the receiver controls for comfortable copying.

Now turn S_2 to the "clipper" position. In order to become familiar with the action of the clipper these steps should be followed: Adjust the "clipping" control so no clipping occurs (maximum positive bias on the diode plates). Set the "clip level" control on the unit so that there will be no apparent change in the strength of the c.w. signal when switching from "clipper" to "out" and back to "clipper." Then turn the "clipping" control until the positive bias is low enough to cause limiting to start; the point at which limiting begins can be recognized by the fact that the signal strength begins to decrease. Back off slightly with the "clipping" control so that the signal strength in the phones is just at the original level.

Tuning the receiver without the use of the limiter shows signals of all strengths, some so loud as to be ear-breaking; but switching to "clipper" will make these big ones drop down to the "comfortable" preset level.

The filter can be aligned with the help of an audio signal generator and a scope. The procedure is to set the two tuned circuits individually to within 10 to 15 cycles of the chosen peak frequency, but on opposite sides of that frequency. This adjustment can be made by tightening or loosening the clamping screws on each choke until each circuit is tuned to the desired frequency. Altering the number of layers of paper placed between the "I" and "E" laminations of either or both chokes will allow any two similar chokes which, due to manufacturing tolerances, may be of slightly different inductances, to be tuned to the same frequency. The filter is then ready to go. If the response is too sharp, slightly greater separation of the two frequencies can be achieved by readjusting the clamp on one of the chokes.

In order to peak a desired signal the receiver b.f.o. or tuning control should be adjusted so the pitch of the signal is 700 cycles. Since the selectivity curve is rather sharp, any adjacent undesired signals will fall short of the peak and be attenuated. If the receiver b.f.o. has sufficient range to tune 700 cycles or more on both sides of zero beat, the undesired signal can always be placed on the notch side of the peak.

A Simple Audio Limiter

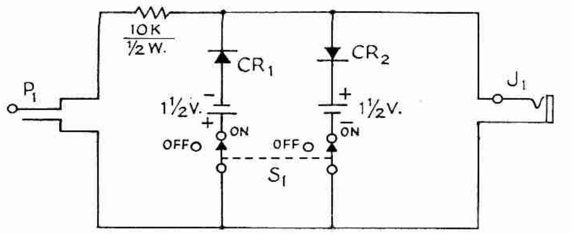


Fig. 5-53—Circuit diagram of a simple audio limiter.
CR_1, CR_2—1N34A or similar germanium diode.
J_1—Open-circuit headphone jack.
P_1—Headphone plug.
S_1—D.p.s.t. toggle or rotary switch.
A Keystone battery holder No. 155 (Allied Radio) will hold two Burgess N, Eveready W468 or Ray-o-Vac 716 flashlight cells.

A simple audio limiter to hold down static crashes and key clicks can be made from two flashlight cells, two germanium diodes and a few other parts. Its use requires no alteration of the receiver, since it is plugged in at the output jack of the receiver and the headphones are plugged into the limiter. A suitable circuit is shown in Fig. 5-53. No constructional details are given because there is nothing critical. If desired, the parts can be housed in a small utility cabinet or "Minibox." Leads can be soldered directly to the flashlight cells or, if desired, a suitable battery holder can be obtained from a radio or model airplane store. Hold the germanium diode leads with pliers when soldering, to prevent heat from reaching and injuring the crystals.

DCS-500 Double-Conversion Superheterodyne

The receiver shown in Fig. 5-54 was designed to meet a need for a better-than-average ham receiver *requiring* a minimum of mechanical work and using standard and easily obtainable parts. It incorporates such features as a 100-kc. calibrator, provision for reception on all ham bands from 80 through 10 meters, adequate selectivity for today's crowded bands, and stability high enough for copying s.s.b. signals. Dubbed the DCS-500 because of its 500-cycle selectivity in the sharpest i.f. position, it is a double-conversion superheterodyne receiver capable of giving good results on either a.m., c.w. or s.s.b.

The Circuit

Referring to the circuit diagram, Figs. 5-55 and 5-56, a 6BA6 r.f., stage is followed by a 6U8A mixer-oscillator. The 4.5-Mc. mixer output is amplified by a 6BA6 and filtered by a two-stage crystal filter, after which a 6U8A second mixer-oscillator, crystal-controlled, heterodynes the signal to 50 kc.

The combination of i.f. amplifiers may appear rather unusual at first glance, since one might expect that a cascade crystal filter in the high-frequency i.f. would make further selectivity unnecessary. This would be true with highly developed filters, but two filters are needed if the best possible job is to be done on both phone and c.w., and such filters are expensive. With inexpensive surplus crystals such as are used in this receiver it would be difficult, if not impossible, to match the performance of the high-class filters; in addition, special test equipment and extreme care in adjustment would be necessary. The approach used here is to use the surplus crystals without such special adjustment, thereby achieving a good, if not quite optimum, degree of selectivity against strong signals near the desired one, and then to back up the filter by a low-frequency i.f. amplifier that will give the "close-in" straight-sided selectivity needed in present-day operation. The overall result is a high order of protection against strong interfering signals at considerably less cost, for the entire double-i.f. system, than that of two high-performance filters alone. The choice of 4.5 Mc., approximately, for the first i.f. was based on the availability of surplus crystals around this frequency, with due consideration for minimizing spurious responses. A second i.f. of 50 kc. was chosen because it lent itself nicely to the utilization of low-cost TV horizontal-oscillator coils as i.f. transformers.

The two i.f. amplifiers at 50 kc. contribute the necessary adjacent-channel selectivity. Three degrees of selectivity are available, depending on the degree of capacitive coupling between the two windings of each i.f. transformer. The greater the number of capacitors switched in parallel — that is, the larger the coupling capacitance — the lower the coupling between the windings and thus the greater the selectivity.

A standard diode detector develops the audio output for all reception modes. The output of the detector is simultaneously applied to both the first audio amplifier and the audio a.g.c. circuit. A series-type noise limiter can be used on a.m. to reduce impulse-noise interference, but this type is ineffective on c.w. or s.s.b. because of the large amplitude of the b.f.o. injection voltage.

The b.f.o., a Hartley-type oscillator, can be tuned from 3 kc. above to 3 kc. below its 50-kc. center frequency by the tuning capacitor.

The first audio stage is a normal Class A voltage amplifier with its output either coupled to the grid circuit of the audio output tube or to a phone jack. High-impedance head-phones (20,000 ohms a.c. impedance or higher) are required. Plugging in the phones automatically disconnects the speaker. If low-impedance headphones are used, they can be connected to the speaker terminals. Capacitances shunting the grid resistors restrict the audio response to an upper limit of about 4000 cycles.

The audio output transformer couples to a low-impedance (3.2-ohm) speaker. The 47-ohm resistor across the secondary protects the transformer in the absence of a speaker load.

The audio output of the detector is also amplified separately in the audio a.g.c. circuit and then rectified to develop a negative voltage that can be used for a.g.c. on c.w. and s.s.b. Two different time constants are used in the rectifier filter circuit, for either fast- or slow-decay a.g.c.

The 100-kc. calibrator employs two 2N107 p-n-p transistors, one as the oscillator and the second as a 100-kc. amplifier. Its transistors obtain the necessary operating potential from the cathode resistor of the audio output tube. Output from the 100-kc. unit is capacity-coupled to the antenna winding of the r.f. coil. Calibrating signals at 100-kc. intervals are *available on all* frequencies covered by the receiver.

The calibrator unit is constructed in a separate

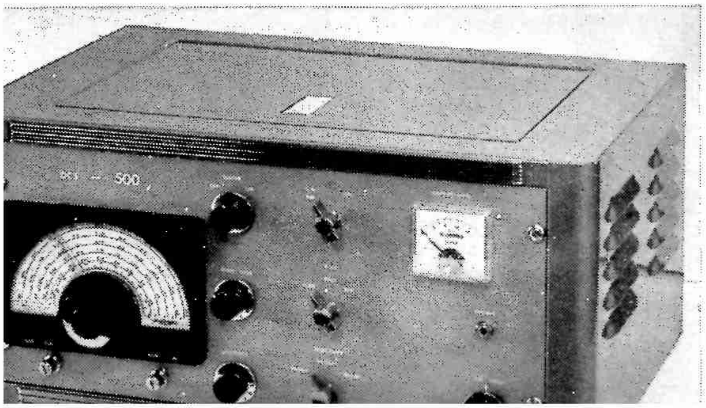


Fig. 5-54—The DCS-500 double-conversion superheterodyne. Left bottom, antenna trimmer, 100-kc. calibrator switch; center, left, top to bottom, noise-limiter switch, volume control, sensitivity control; center, right, b.f.o. switch, a.g.c. speed, selectivity; right, headphone jack, b.f.o. pitch control. The dial is a National ICN. Front panel is 8¾ inches high; the receiver is mounted in a Bud CR-1741 rack cabinet.

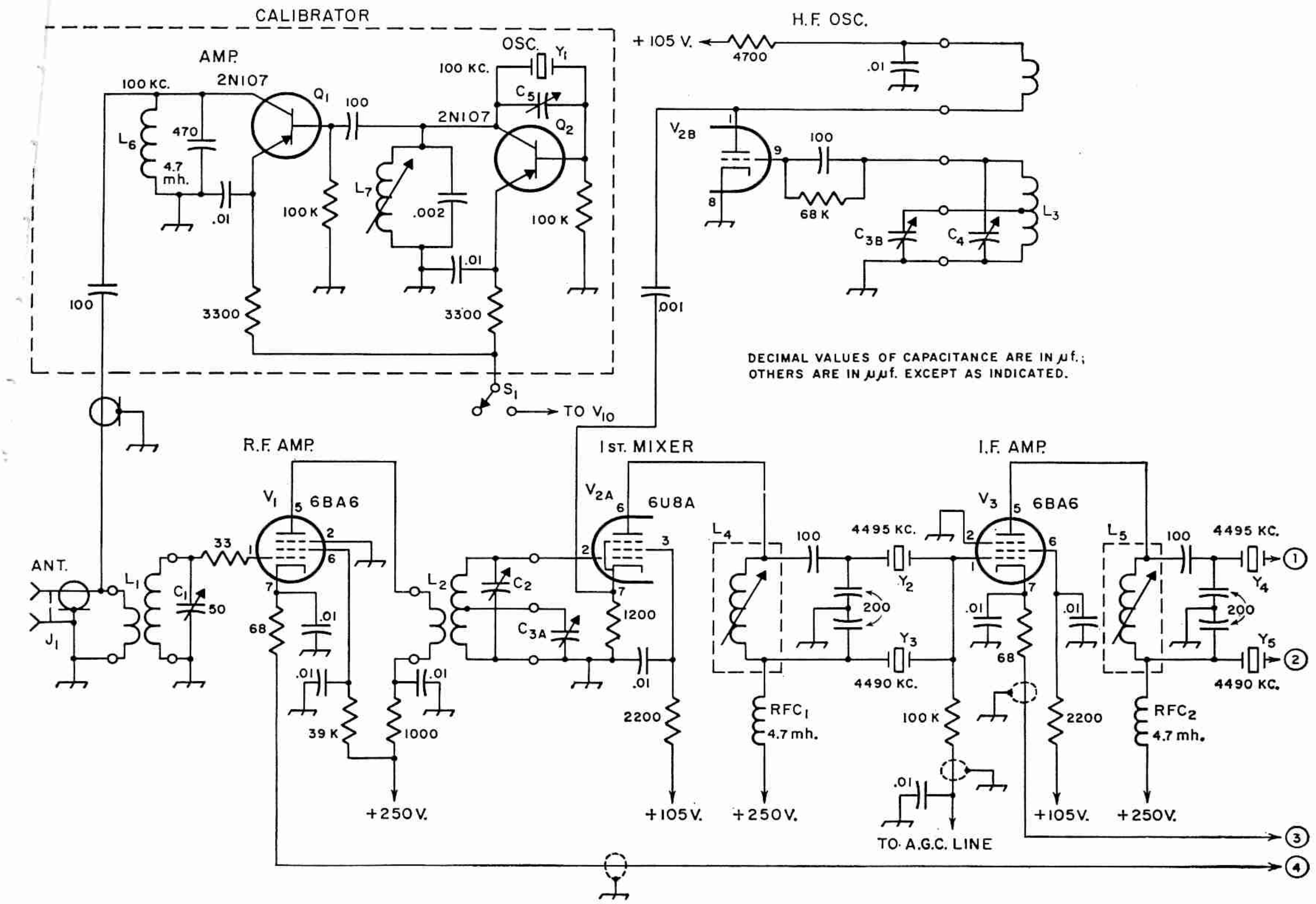


Fig. 5-55—Front-end circuit of the receiver. Unless otherwise specified, resistors are ½ watt; 0.01 and 0.02-µf. capacitors are disk ceramic, 600 volts; 0.5 capacitors are tubular paper; capacitors below 0.01 µf. are mica; capacitors marked with polarities are electrolytic.

C_1—50-µµf. variable (Hammarlund HF-50).
C_2, C_4—See coil table.
C_3—2-section variable, 5—28.5 µµf. per section, double spaced (Hammarlund HFD-30-X).
C_5—3-30-µµf. ceramic trimmer.
J_1—Coaxial receptacle, chassis mounting (SO-239).
L_1, L_2, L_3—See coil table.
L_4, L_5—18-36-µh. slug-tuned (North Hills 120E coil mounted in North Hills S-120 shield can).
L_6—4.7 mh. (Waters C1061).
L_7—1-2-mh. slug-tuned (North Hills 120K).
RFC_1, RFC_2—4.7 mh. (Waters C1061).
S_1—Single-pole rotary.
Y_1—100 kc. (James Knights H-93).
Y_2, Y_4—4495 kc. (surplus).
Y_3, Y_5—4490 kc. (surplus).

Minibox so that it can be plugged into the accessory socket of the receiver or used as an individual unit powered by penlite cells.

The power supply, Fig. 5-57, is a full-wave rectifier with a choke-input filter. It provides approximately 250 volts d.c. under load. A 0.25-µf. capacitor is shunted across the 10-henry filter choke to form a parallel-resonant circuit at 120 cycles; this provides an increased impedance to the ripple component and thus reduces hum in the output of the supply.

The power-supply requirements are 250 volts at 110 milliamperes, and 6.3 volts at approximately 5 amperes. Any transformer-choke combination fulfilling the requirements can be used.

Front End

The use of plug-in coils for the front end eliminated the mechanical problems of a band-switching tuner, and also offered the possibility of realizing higher-*Q* tuned circuits. Ganged tuning of the r.f. amplifier along with the h.f. oscillator and mixer circuits was decided against because of the complexities it would cause in coil construction and the problem of keeping three stages tracking with each other. The r.f. amplifier has to be peaked separately by the antenna trimmer, but separate peaking insures maximum performance at all frequencies.

Construction

The receiver is constructed on a 12 × 17 × 2-inch aluminum chassis with an 8¾ × 19-inch aluminum front panel, which permits it to be installed in a table-type rack cabinet. The general layout of components can be seen in Figs. 5-58 and 5-60. A good procedure to follow when

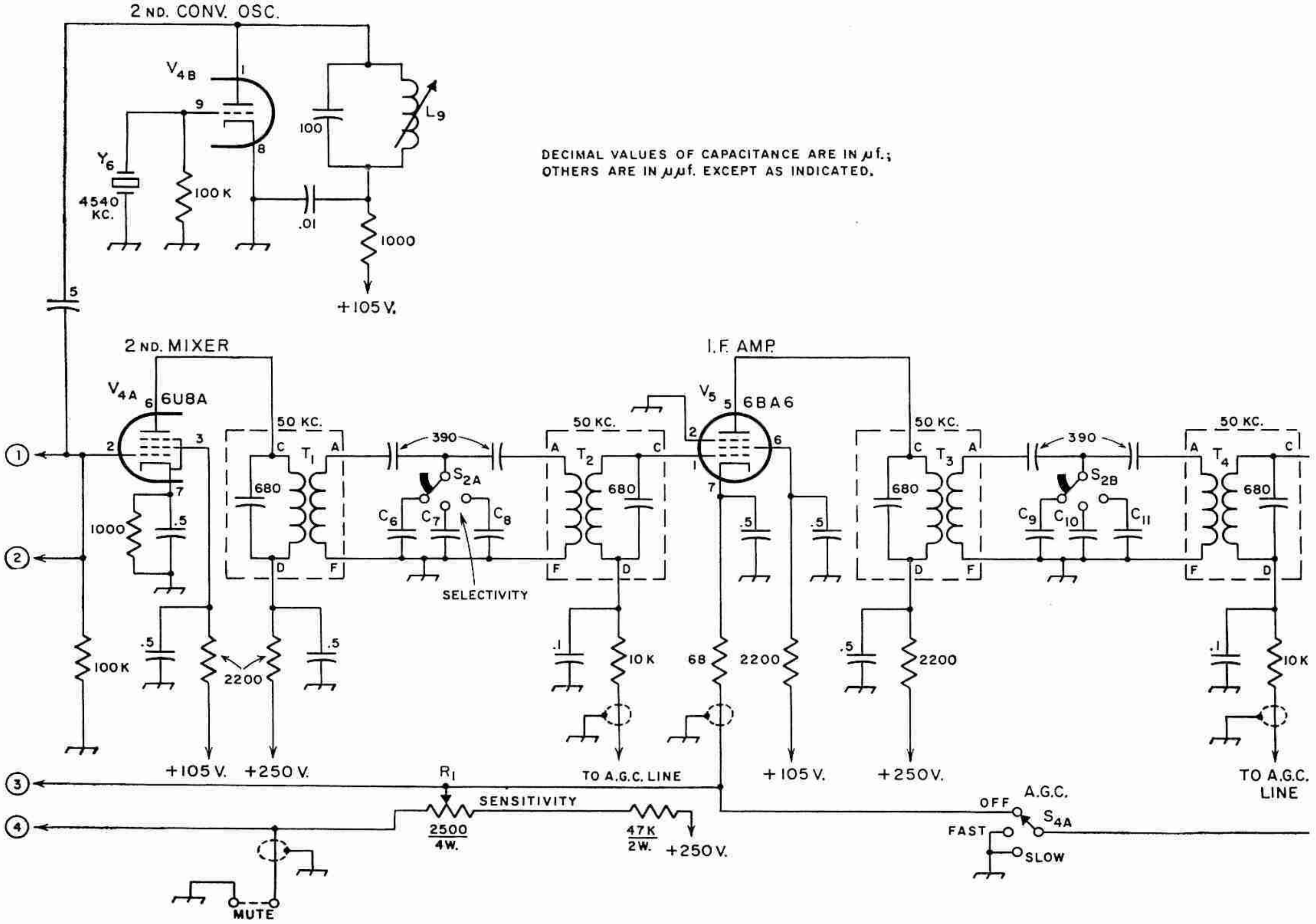


Fig. 5-56—I.f. amplifier, detector, a.g.c. and audio circuits. Unless otherwise specified, resistors are ½ watt; 0.01- and 0.02-μf. capacitors are disk ceramic, 600 volts; 0.5-μf. capacitors are tubular paper; capacitors below 0.01 μf. are mica; capacitors marked with polarities are electrolytic.

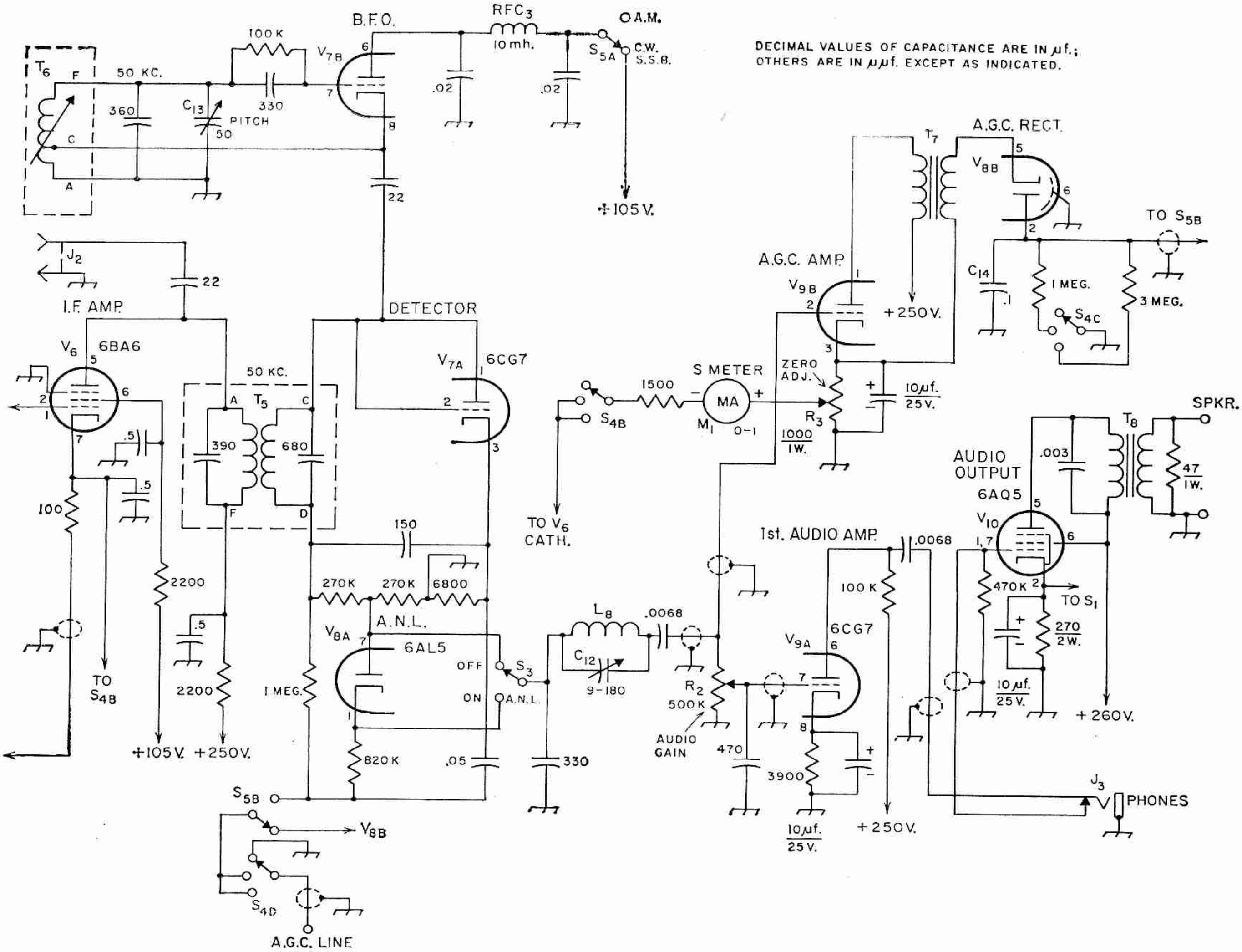
C_6, C_7, C_8, C_9, C_{10}, C_{11}—0.01 mica (Aerovox CM-30B-103)
C_{12}—9-180-μμf. mica compression trimmer.
C_{13}—50-μμf. variable (Hammarlund HF-50).
C_{14}—0.1-μf. paper (Sprague 2TM-P1).
J_2—Phono jack.
J_3—Closed-circuit phone jack.
L_8—125 mh. (Meissner 19-6848).
L_9—9-18 μh., slug-tuned (North Hills 120D).
M_1—0-1 d.c. milliammeter (Triplett 227-PL).
R_1—2500-ohm, 4-watt control, wire-wound.
R_2—0.5-megohm control, audio taper with push-pull type switch (S_6) (Mallory No. PP55DT1683).
R_3—1000-ohm, 1-watt control, wire-wound.
RFC$_3$—10 mh. (National R-50-1).
S_1, S_3—Rotary, 1 section, 1 pole, 2 position.
S_2—Rotary, 2 section, 1 pole per section, progressively shorting. Switch section Centralab PA-12, index Centralab PA-302.
S_4—Rotary, 1 section, 5 poles per section (4 poles used), 3 positions used, Centralab PA-2015.
S_5—Rotary, 1 section, 2 poles per section, 2 positions used. Centralab PA-2003.
T_1–T_5, inc.—50-kc. i.f. transformers made from TV components (Miller 6183); see text.
T_6—B.f.o. transformer (Miller 6183); see text.
T_7—Audio interstage transformer, 1:2 ratio (Thordarson 20A16).
T_8—Audio output transformer, 5000 to 4 ohms (Stancor 3856).
Y_6—4540 Kc. (surplus).

starting to wire the receiver is first to complete the power supply and heater wiring, and then start wiring from the antenna toward the speaker. This allows proceeding in a logical order so that the work can be picked up readily at any time after an intermission.

The use of good quality ceramic tube and coil sockets, particularly in the front end, is highly recommended. When mounting the sockets orient them so that the leads to the various points in the circuit will be as short as possible.

Millen coil shields (80008) are used around the plug-in coils in the front end — i.e., the r.f., mixer and oscillator — and the shield bases are mounted with the same screws that hold the ceramic coil sockets. All plug-in coils are wound with No. 26 enameled wire on Amphenol polystyrene forms, and Hammarlund APC-type airpadder capacitors are mounted in the recesses at the tops of the coil forms. After finishing a coil it is a good idea to fasten the winding and the trimmer capacitor in place with Duco cement. Decal each set of coils for a particular band and mount them on small wooden bases that have holes to take the pins Then paint or stain each of the coil-set bases. The final result will be a neat and convenient arrangement for holding the coils for each band (Fig. 5-59). Plug-in coil data for each band are given in the coil table.

The tuning capacitor, C_3, is mounted on the



chassis and reinforced by a bracket to minimize any rocking movement. This bracket is triangular in shape with a right-angle flange at the bottom. It is drilled to take the front bearing sleeve of the tuning capacitor and held firmly to it by the capacitor mounting nut and a lock washer, as shown in Fig. 5-58. Flexing of the chassis can be minimized by the use of lengths of angle stock bolted to the chassis at strategic points throughout the receiver. Exact alignment of the tuning capacitor with the dial shaft is not always possible, so a flexible coupling (Millen 39016) is used.

When wiring the crystal filter keep leads as short and direct as possible, as this will minimize stray coupling between the input and output ends, which would deteriorate the performance of the crystal-filter circuits.

The 50-kc. i.f. circuits used Miller 6183 TV horizontal-oscillator replacement coils as i.f. transformers. These coils must be altered before they can be used. As they are supplied, the terminal lugs are brought out at the top of the can; these lugs must be reversed before the can is mounted. By applying slight pressure to the phenolic coil form the assembly will slide out of the aluminum shield can and then can be reversed. However, before reassembling the unit a few slight changes must be made. There are actually two separate windings; each one will be tuned and used either as a primary or secondary for the 50-kc. i.f. transformer. The tap on the large winding must be lifted off the soldering lug C, taped, and tucked away, being careful not to break it; this leaves just the lead from the small winding on terminal C. Terminals A and F represent the large winding. The small coil is tuned by connecting a 680-µµf. mica capacitor between terminals C and D; these capacitors should be fastened on the soldering lugs inside the shield can. The can is then slipped back over the coil and capacitor, keeping in mind that the lugs must come out the bottom, and the assembly is ready for mounting on the chassis.

The b.f.o. coil is also a Miller 6183, and the procedure for reversing the assembly before mounting is identical to that followed with the 50-kc. transformers. However, it is not necessary to alter any of the wiring in the b.f.o. transformer, since only the large winding (A — F) and its tap (C) is used.

Point-to-point wiring is recommended, along with generous use of both insulated tie points and

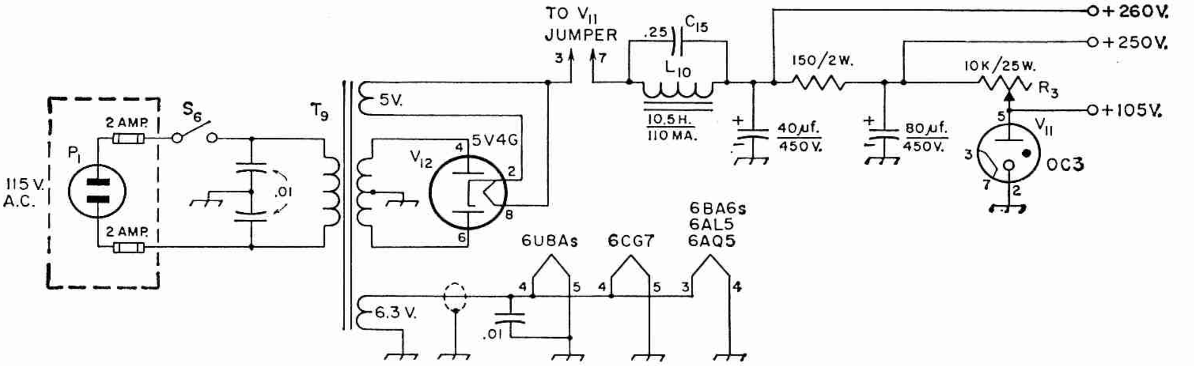


Fig. 5-57—Power-supply circuit. Capacitors marked with polarities are electrolytic.

C_{15}—0.25-µf. paper, 600 volts.
L_{10}—Filter choke, 10.5 henry, 110 ma. (Knight 62 G 139).
P_1—Fuse plug.
S_6—See R_2, Fig. 5-56.
T_9—Power transformer, 700 v. c.t., 120 ma.; 5 v., 3 amp.; 6.3 v., 4.7 amp. (Knight 62 G 044).

ground lugs. Use of shielded wire facilitates routing wires throughout the receiver as the shields can be spot-soldered to ground lugs and to each other in bundles. When wiring, mount components at right angles to the chassis sides wherever possible; this helps give the finished unit a neat appearance. In critical circuits, however, do not sacrifice short and direct leads for the sake of making the unit look pretty.

Placing the receiver in a rack cabinet and marking all controls on the front panel with decals also helps in giving the finished receiver a neat and "commercial" appearance.

The Calibrator

The 100-kc. calibrator is built in a separate 4 × 4 × 2-inch aluminum box and plugs into the accessory socket at the left rear of the receiver chassis. Fig. 5-61 shows the internal construction. The accessory socket provides the necessary operating voltage for the transistors and offers a convenient means for coupling the 100-kc. harmonics out of the calibrator into the receiver. If the calibrator is to be used as a self-contained unit it must be supplied with approximately 7–10 volts. A series arrangement of penlite cells, or a mercury battery, can be used. A battery clip mounted on the side of the box is a convenient way to hold the internal batteries. If the unit is to be self-contained, a separate output jack for the calibrator must be provided. A phono jack may be used.

I.F. Alignment

Before starting alignment of the receiver, first determine whether the audio stages are functioning correctly. An audio signal should be coupled to the top end of the volume control, and varying the control should change the output level of the audio signal. If an audio signal is not available, the 60-cycle heater voltage will provide a convenient audio signal for checking.

There are various ways to approach the alignment problem. A 50-kc. signal generator can be used; however, these are hard to come by. Some of the better audio oscillators go as high as 50 kc. and can be used for alignment purposes. A second, and possibly superior, method is to use any of the numerous signal generators which will deliver 4.5-Mc. output; fed into the first i.f. amplifier grid, the 4.5-Mc. signal will beat against the second conversion oscillator to produce a 50-kc. i.f. signal which then can be used for alignment. This method also insures that the first i.f. signal

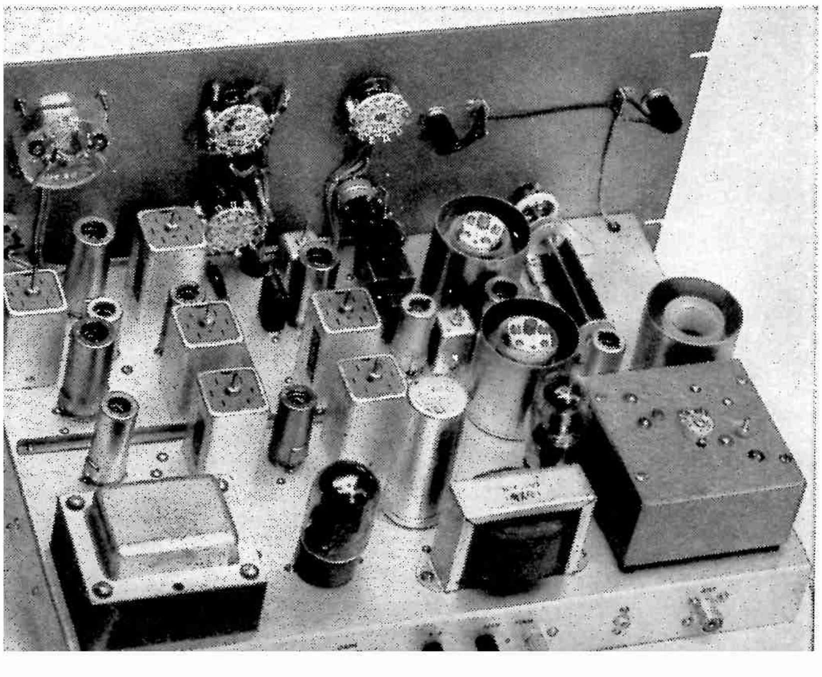


Fig. 5-58—The power supply is built along the back of the chassis; filter capacitor and VR tube are just in back of the filter choke in this view. The crystal calibrator unit at right is cushioned by rubber bumpers mounted on the receiver chassis. C_5 is on top of the calibrator unit. Front-end coil shields are at the top right in this photograph, along with the tuning capacitor bracket and flexible coupling. The on-off switch, S_6, on rear of the audio gain control, is a new push-pull type. Filter crystals are grouped behind the volume control, and the second conversion oscillator crystal is slightly to their left. The 4.5-Mc. i.f. transformers (in the small shield cans) are close to the filter crystals. The b.f.o. coil is at the extreme left in this view; all other large cans contain the 50-kc. i.f. transformers. Connections on the back chassis wall, left to right, are the muting terminals, B-plus output for auxiliary use, speaker terminals, i.f. output (phono jack), and antenna input connector.

Fig. 5-59—Each set of coils is provided with a wooden base for storage. C_2 and C_4 are mounted in the recesses at the tops of the oscillator and mixer coil forms.

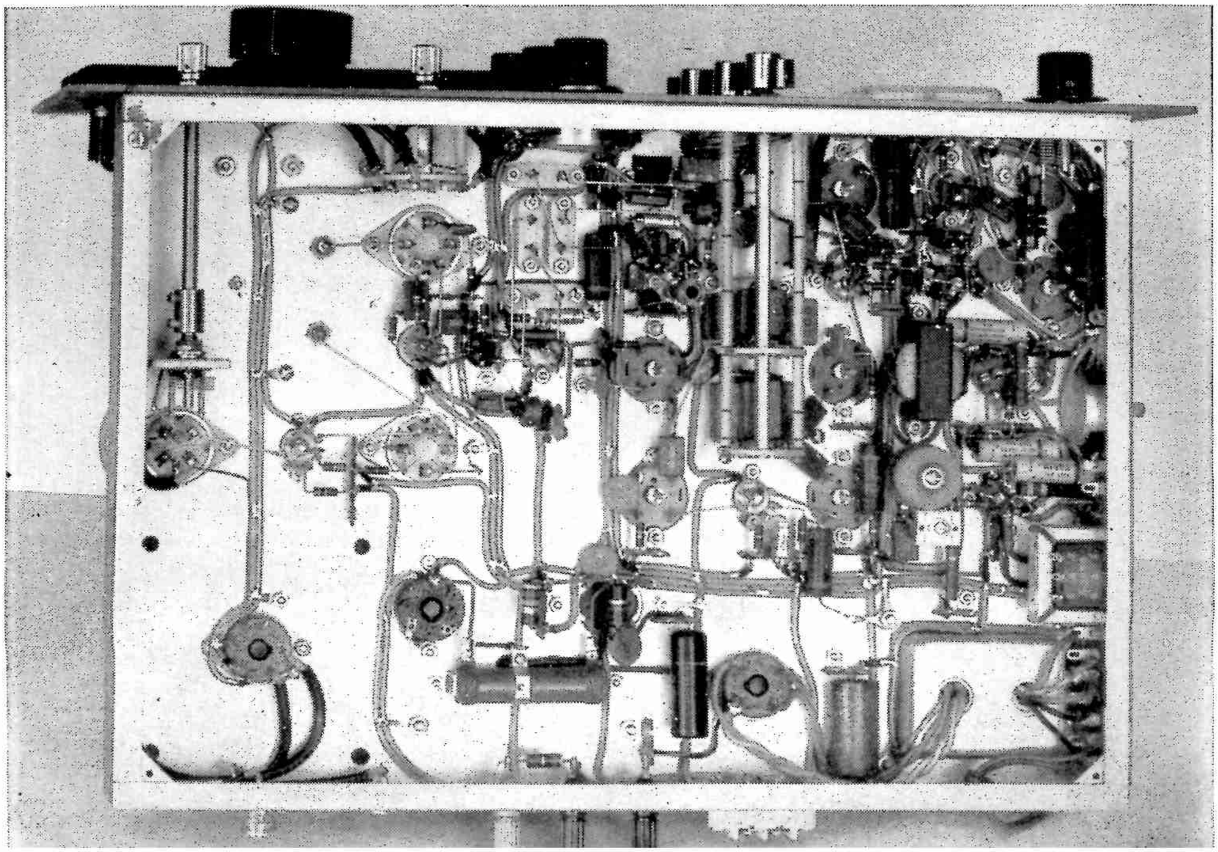
DCS-500 Coil Table

All coils wound with No. 26 enameled wire on $1\frac{1}{4}$-inch diameter polystyrene forms. R.f. coil forms are four-prong (Amphenol 24-4P); mixer and oscillator coils are five-prong (Amphenol 24-5P). C_2 and C_4 are Hammarlund APC-50 except on 3.5 Mc., which takes APC-75. Taps are counted from ground end. Primaries and ticklers are close-wound in the same direction as the main coil at bottom of coil form; grid and plate (or antenna) connections at outside ends.

Band	*Secondary*	*Primary or Tickler*
3.5 Mc.	L_1, $45\frac{1}{4}$ turns close-wound.	$10\frac{3}{4}$ turns, $\frac{3}{8}$-inch spacing from secondary.
	L_2, $36\frac{3}{4}$ turns close-wound, tapped at $26\frac{3}{4}$ turns.	$11\frac{3}{4}$ turns, $\frac{1}{4}$-inch spacing from secondary.
	L_3, $28\frac{3}{4}$ turns close-wound, tapped at 19 turns.	$7\frac{3}{4}$ turns, $\frac{1}{4}$-inch spacing from secondary.
7 Mc.	L_1, $26\frac{1}{4}$ turns, close-wound.	$7\frac{3}{4}$ turns, $\frac{1}{4}$-inch spacing from secondary.
	L_2, $18\frac{3}{4}$ turns spaced to 1 inch. Tapped at $9\frac{3}{4}$ turns.	$6\frac{3}{4}$ turns, $\frac{3}{8}$-inch spacing from secondary.
	L_3, $17\frac{3}{4}$ turns spaced to $\frac{7}{8}$ inch. Tapped at $4\frac{3}{4}$ turns.	$7\frac{3}{4}$ turns, $\frac{1}{4}$-inch spacing from secondary.
14 Mc.	L_1, $13\frac{1}{4}$ turns spaced to $\frac{5}{8}$ inch.	$6\frac{3}{4}$ turns, $\frac{5}{16}$-inch spacing from secondary.
	L_2, $10\frac{3}{4}$ turns spaced to 1 inch. Tapped at 3 turns.	$5\frac{3}{4}$ turns, $\frac{5}{8}$-inch spacing from secondary.
	L_3, $5\frac{3}{4}$ turns spaced to $\frac{15}{16}$ inch. Tapped at $1\frac{7}{8}$ turns.	$3\frac{3}{4}$ turns, $\frac{5}{8}$-inch spacing from secondary.
21 Mc.	L_1, $9\frac{1}{4}$ turns spaced to $\frac{1}{2}$ inch.	$6\frac{3}{4}$ turns, $\frac{5}{16}$-inch spacing from secondary.
	L_2, $7\frac{3}{4}$ turns spaced to $1\frac{1}{8}$ inches. Tapped at 2 turns.	$5\frac{3}{4}$ turns, $\frac{5}{8}$-inch spacing from secondary.
	L_3, $6\frac{3}{4}$ turns spaced to $\frac{15}{16}$ inch. Tapped at 2 turns.	$3\frac{3}{4}$ turns, $\frac{3}{8}$-inch spacing from secondary.
28 Mc.	L_1, $6\frac{1}{4}$ turns spaced to $\frac{7}{16}$-inch.	$5\frac{3}{4}$ turns, $\frac{1}{4}$-inch spacing from secondary.
	L_2, $5\frac{3}{4}$ turns spaced to $1\frac{1}{4}$ inches. Tapped at 2 turns.	$4\frac{3}{4}$ turns, $\frac{1}{4}$-inch spacing from secondary.
	L_3, $4\frac{3}{4}$ turns spaced to $\frac{3}{4}$ inch. Tapped at $1\frac{1}{2}$ turns.	$2\frac{3}{4}$ turns, $\frac{1}{4}$-inch spacing from secondary.

will fall within the crystal filter bandpass in case the crystal frequencies are not exact. When aligning, connect a d.c. voltmeter (preferably a v.t.v.m.) across the detector load resistor (point D of T_5 and chassis), turn the i.f. gain control about three-quarters open, and tune both the plate circuit of the second conversion oscillator and the 50-kc. i.f. transformers for maximum output, as indicated on the meter. The output of the signal generator should not be modulated, and at the start will most likely be "wide open." However, as alignment progresses the output of the generator will have to be progressively decreased. When aligning the i.f. transformers there should

Fig. 5-60—The potentiometer for S-meter adjustment and the audio output transformer are on the right chassis wall in this view. The 50-kc. i.f. trap is located just above the power transformer in the lower right-hand corner. The antenna trimmer is located at extreme left center. The crystal filter sockets are at top center, and to their left on the front wall is the calibrator switch S_1. To the right of the calibrator switch is the sensitivity control, followed to the right by the selectivity switch S_2 and the b.f.o. pitch-control capacitor. The octal accessory socket for the calibrator is at the lower left. As shown, shielded wire spot-soldered together in bundles can be routed conveniently to various points in the receiver. Ceramic sockets are used throughout the front end (center left). Mounting components parallel with the chassis sides helps give the finished unit a neat appearance.



be a definite peak in output as each circuit is brought through resonance. If a particular coil does not peak, that coil and its associated circuits should be checked. After peaking one winding of a transformer, recheck the other; it may need touching up. After alignment of all the 50-kc. coils is completed, go back and "rock" each coil slug to be sure it is peaked for maximum output. This completes the 50-kc. alignment.

Leave the signal generator on, set the b.f.o. pitch control at half capacitance, turn the b.f.o. on, and adjust its coil slug for zero beat with the 50-kc. i.f. signal. Varying the pitch control over its range should produce a tone with a maximum frequency of 3 kc. either side of zero beat.

Next, the 50-kc. trap on the output of the detector should be adjusted. Connect the vertical input terminals of an oscilloscope between the plate of the first audio amplifier and chassis, turn on the b.f.o., and adjust C_{12} for minimum 50-kc. signal on the scope. This trap, made up of C_{12} and L_8, attenuates any 50-kc. feed-through.

The first-i.f. coils at 4.5-Mc. should next be adjusted. Couple the signal generator to the grid of the first mixer and peak L_4 and L_5 for maximum deflection of the v.t.v.m. at the detector. The i.f. system is then completely aligned.

Front-End Alignment

To adjust the front end, plug in a set of coils and check the oscillator frequency range either with a calibrated g.d.o. or on a calibrated general-coverage receiver, the latter being preferable. Keep in mind that the oscillator works 4.5 Mc. above the signal on 80, 40 and 20 meters, and 4.5 Mc. below the signal frequency on the 15- and 10-meter bands. This means that on 15 and 10 meters the oscillator trimmer capacitor, C_4, must be at the larger-capacitance setting of the two that bring in signals. After establishing the correct frequency range of the oscillator, inject a signal at the low end of the band into the antenna terminals and peak the mixer capacitor, C_2, and the antenna trimmer, C_1, for maximum signal. Then move the test signal to the high end of the band and recheck the mixer trimmer capacitor (the antenna trimmer also will have to be repeaked) for correct tracking. If C_2 has to be readjusted, spread the mixer coil turns apart or compress them together until the signal strength is uniform at both ends of the band, without readjustment of C_2. If the mixer trimmer capacitance had to be increased at the high-frequency end of the band to maintain tracking, the coil tap is too far up the coil and the turns below the tap must be spread apart or the tap itself must be moved down. If the trimmer capacitance has to be decreased the tap is too low. Coil specifications might possibly have to be altered slightly from those given in the table, particularly on the higher frequencies, because of variations in strays from one receiver to another.

General

Adjustment of the calibrator is relatively straightforward, and should present no problems. Turn on the calibrator and you should hear the 100-kc. harmonics on whatever band you happen to be using. Once it is determined that the unit is working correctly, the only adjustment necessary is to set the frequency of the calibrator exactly. The usual reference is WWV or any broadcast station that is on a frequency which is a whole-number multiple of 100 kc. The frequency tolerance for standard broadcast stations is 20 cycles, thus b.c. stations represent a source for accurate frequency determination.

Using a general-coverage or b.c. receiver, tune in either WWV or a known broadcast station and adjust the calibrator trimmer C_5 for zero beat. The calibrator will then provide accurate 100-kc. signals that can be used for frequency determination and band-edge marking.

The first intermediate frequency can be altered slightly to facilitate the use of particular sets of crystals available. However, if the deviation is more than 20 kc. or so, slight changes may be needed in the h.f. oscillator coil specifications to maintain the proper bandspread.

If the receiver is to be worked in a rack cabinet as shown in Fig. 5-54, or if a cover plate is attached to the bottom of the receiver chassis, minor alignment touch-up may be necessary.

Spraying the receiver chassis with a light coat of clear plastic lacquer before mounting any of the components will prevent fingerprints and oxidation of the chassis.

The audio output stage has adequate power to drive a 5- or 6-inch speaker, which may be mounted in a small open-back metal utility box.

The i.f. output jack at the rear provides a convenient way of attaching accessory devices such as an oscilloscope for modulation checking.

A side-by-side comparison of the finished receiver with some of the better-quality commercial units will show that this receiver can hold its own in sensitivity, selectivity and stability. Needless to say, the more care taken in construction, wiring and alignment the better the results.

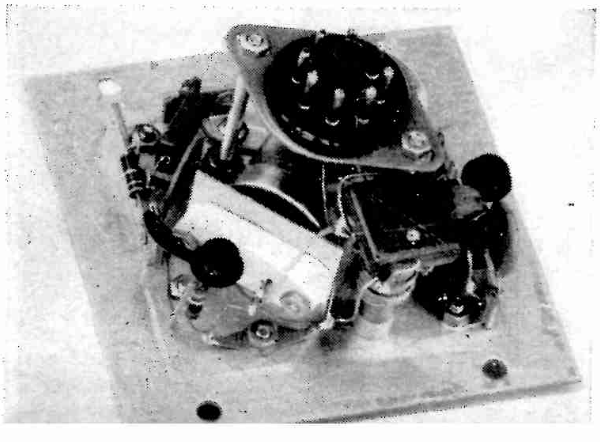


Fig. 5-61—Inside view of the calibrator unit. The 100-kc. oscillator coil, L_{15}, is at the right, the oscillator transistor, Q_2, is in the foreground mounted to the crystal socket, and the amplifier transistor, Q_1, is mounted at the right on a terminal strip. The 100-kc. crystal is mounted horizontally between the plate and the octal plug. The plug can be mounted on 2-inch screws as shown in the photograph, or on the bottom plate of the Minibox, with flexible leads to the circuit. If the calibrator is to be used as a self-contained unit (see text) the octal plug is not necessary.

A Transistorized Q Multiplier

A "*Q* multiplier" is an electronic device that boosts the *Q* of a tuned circuit many times beyond its normal value. In this condition the single tuned circuit has much greater selectivity than normal, and it can be utilized to reject or amplify a narrow band of frequencies. There are vacuum-tube versions of the *Q*-multiplier circuit, but the transistorized *Q* multiplier to be described has the advantage that it eliminates a power-supply problem and is very compact.

Circuit and Theory

Parallel-tuned circuits have been used for years as "suck-out" trap circuits. Properly coupling a parallel-tuned circuit loosely to a vacuum-tube amplifier stage, it will be found that the amplifier stage has no gain at the frequency to which the trap circuit is tuned. The additional tuned circuit puts a "notch" in the response of the amplifier. The principle is used in TV and other amplifiers to minimize response to a narrow band of frequencies. Increasing the *Q* of the trap circuit reduces the width of the rejection notch.

The transistorized *Q* multiplier makes use of the above effect for its operation. A tuned circuit is made regenerative to increase its *Q* and is coupled into the i.f. stage of a receiver. By changing the frequency of the regenerative circuit, the sharp notch can be moved about across the passband of the receiver. The width of the notch is changed by controlling the amount of regeneration.

Although it seems paradoxical, the transistorized *Q* multiplier with no change in circuitry will also permit "peaking" an incoming signal the way a vacuum-tube *Q* multiplier does. The mode of operation is selected by adjustment of the regeneration control, and this then usually requires a slight readjustment of the frequency control. The peaking effect is not quite as pronounced as the notch, but it is still adequate to give fairly good single-signal c.w. reception with a receiver of otherwise inadequate selectivity.

The regenerative circuit builds up the signal and feeds it back to the amplifier at a higher level and in the proper phase to add to the original signal. The notch effect described earlier works in a similar manner except that the tuning of the regenerative circuit is such that it feeds back the signal out of phase.

The schematic diagram of the *Q* multiplier is shown in Fig. 5-62. The inductor L_1 furnishes coupling from the receiver to the *Q* multiplier, and C_4 is required to prevent short-circuiting the receiver's plate supply. The multiplier proper consists of the tunable circuit $C_1C_3L_2$ connected to a transistor in the collector-tuned common-base oscillator circuit using capacitive feedback via C_2. Regeneration is controlled by varying the d.c. operating voltage through dropping resistor R_1.

Layout

The unit and power supply are built in a small aluminum "Minibox" measuring 5 × 2¼ × 2¼ inches (Bud CU-3004) and the operating controls are mounted on a lucite or aluminum subpanel. All parts of the unit are built on one half of the box. This feature not only simplifies construction but makes a battery change a simple job, even if this is required only a couple of times a year.

All major components, such as the two slug-tuned coils, tie point, battery holder, regeneration and tuning controls, are mounted directly on the box and subpanel. The remaining resistors, capacitors and the single transistor are supported by their connections to the above parts.

The two slug-tuned coils, L_1 and L_2, are centered on the box and spaced one inch apart on centers. Operating controls C_1 and R_1 are placed 1¼ inches from the ends of the subpanel and centered. The tie point mounts directly behind tuning control C_1.

Power for the unit is supplied by four penlight cells (type 912) which are mounted in the battery holder (Lafayette Radio Co. Stock No. MS-170) directly behind regeneration control R_1. Total drain on the battery never exceeds 0.2 ma.

Connection to the receiver is made with a three-foot length of RG-58/U cable brought through the rear wall of the Minibox. A rubber grommet

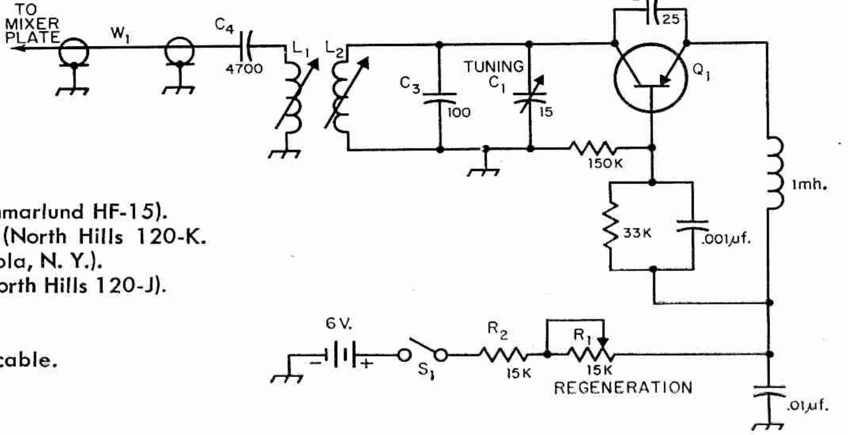


Fig. 5-62—Circuit diagram of the 455-kc. transistorized Q multiplier. Unless otherwise indicated, capacitances are in μμf., resistances are in ohms, resistors are ½ watt.

C_1—15-μμf. variable capacitor (Hammarlund HF-15).
L_1—1000–2000-μh. slug-tuned coil (North Hills 120-K. North Hills Electric Co., Mineola, N. Y.).
L_2—500–1000-μh. slug-tuned coil (North Hills 120-J).
Q_1—CK768 PNP junction transistor.
S_1—Part of R_1.
W_1—Three-foot length of RG-58/U cable.

Fig. 5-63—View of the Q multiplier showing its single connecting cable to the receiver. The box can be placed in any convenient spot on or around the receiver.

should be placed in the hole to prevent chafing of the cable insulation.

When soldering the transistor in place, be sure to take the usual precautions against heat damage.

Alignment

After completing the wiring (and double-checking it) connect the open end of the three-foot cable to the plate circuit of the receiver mixer tube. This can be done in a permanent fashion by soldering the inner conductor of the cable to the plate pin on the tube socket or any point that is connected directly to this pin, and by soldering the shield to any convenient nearby ground point. If you are one of those people who is afraid to take the bottom plate off his receiver, and you have a receiver with octal tubes, a "chicken connection" can be made by removing the mixer tube and wrapping a short piece of small wire around the plate pin. Reinsert the tube in its socket and solder the center conductor of the coax to the small wire coming from the plate pin. Now ground the coax shield to the receiver chassis. It is important to keep the lead from the tube pin to the coax as short as possible, to prevent stray pickup.

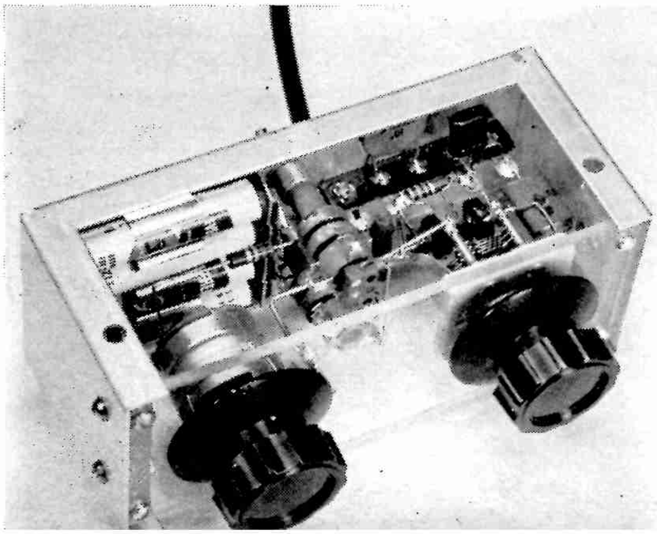


Fig. 5-64—The Q multiplier and its battery supply are combined in one small Minibox. The single transistor is visible near the top right corner.

Check the schematic diagram of the receiver for help in locating the above receiver connections.

Turn on the receiver and tune in a signal strong enough to give an S-meter reading. Any decent signal on the broadcast band will do. Next, tune the slug on L_1 until the signal peaks up. You are tuning out the reactance of the connecting cable, and effectively peaking up the i.f. If the receiver has no S meter, use an a.c. voltmeter across the audio output. When this step has been successfully completed the Q multiplier is properly connected to the receiver and when switched to "off" (S_1 opened) will not affect normal receiver operation.

The next step is to bring the multiplier into oscillation, and to adjust its frequency to a useful range. Set the tuning control to half capacity and advance the regeneration control to about half open. This latter movement also turns the power on. Tune the receiver to a clear spot and set the receiver b.f.o. to the center of the pass-band. Now adjust the slug of L_2. The multiplier should be oscillating, and somewhere in the adjustment of L_2 a beat note will be heard from the receiver. This indicates the frequency of oscillation is somewhere on or near the i.f. Swing this into zero beat with the b.f.o.

Final Adjustment

One of the best ways to make final alignment is to simulate an unwanted heterodyne in the receiver and adjust the Q multiplier for maximum attenuation of the unwanted signal. To do this, tune in a moderately weak signal with the b.f.o. on. A broadcast station received with the antenna disconnected will do. The b.f.o. will beat with the incoming signal, producing an audio tone. Adjust the b.f.o. for a tone of about 1 kc. or so.

Back off on control R_1 until the oscillator becomes regenerative. By alternately adjusting the tuning control, C_1, and the regeneration control, R_1, a point can be found where the audio tone disappears, or at least is attenuated. Some slight retouching of L_2 may have to be done in the above alignment, since the movement of any one control tends to "pull" the others. The optimum situation is to have the tuning control C_1 set at about half capacity when the notch is in the center of the passband.

If you happen to get a super active transistor and the regeneration control does not have the range to stop oscillator action, increase the value of the series resistor R_2. Conversely, if the unit fails to oscillate, reduce the value of R_2.

When making the above adjustments, you should notice that the audio tone can be peaked as well as nulled. If it can not be peaked, a little more practice with the controls should produce this condition. In the unit shown here, the best null was produced with the regeneration control turned only a few degrees. Optimum peak position was obtained with the regeneration control almost at the point of oscillation.

Conelrad

Effective January 2, 1957, the "Conelrad" rules became part of the amateur regulations. Essentially, compliance with the rules consists of monitoring a broadcast station — standard band, f.m. or TV — either continuously or at intervals not exceeding ten minutes, during periods in which the amateur transmitter is in use. On receipt of a Conelrad Alert all transmitting must cease, except as authorized in 12.193 and 12.194 of the FCC regulations.

The existence of an Alert may be determined as outlined in 12.192(b)(3). Operation during hours when local broadcast stations are not on the air will require tuning through the standard broadcast band to determine if operation appears to be normal. The presence of any U. S. broadcast stations on frequencies other than 640 and 1240 kc. indicates normal operation.

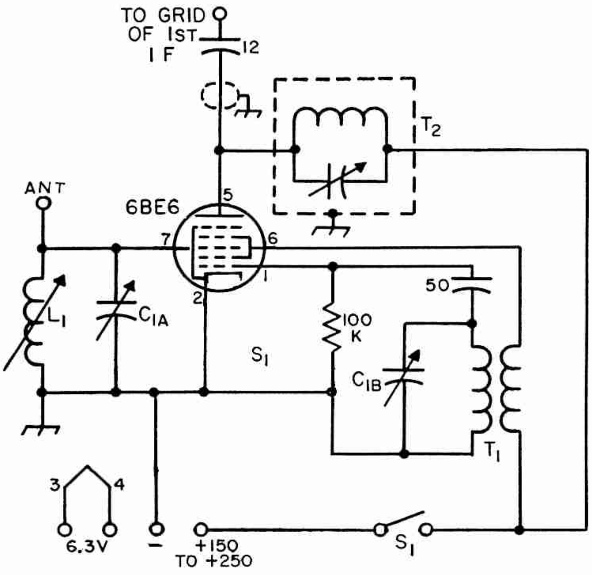


Fig. 5-65—Converter circuit for monitoring broadcast stations in connection with a communications receiver. Capacitances are in $\mu\mu$f.

C_{1A}, C_{1B}—Two-gang broadcast capacitor, oscillator section according to intermediate frequency to be used.

L_1—Loop stick.

T_1—B.c. oscillator transformer (for i.f. to be used).

T_2—I.f. coil and trimmer. This can be taken from an i.f. transformer, or the transformer can be used intact, the output being taken from the secondary.

Note: If only one broadcast station is to be monitored C_{1A} and C_{1B} can be padder-type capacitors (or a combination of padding and fixed capacitance as required) adjusted for the desired station and intermediate frequencies. Other types of converter tubes may be substituted if desired.

Power for the unit can be taken from the receiver's "accessory" socket.

Perhaps the simplest form of compliance is by means of a simple converter working into the i.f. amplifier of the regular station receiver. A typical circuit is shown in Fig. 5-65. The converter can be built in a small metal case and mounted at a convenient spot on the receiver so that S_1 can be closed at regular intervals for checking the broadcast station. As an alternative, the converter can be mounted out of the way at the rear of the receiver and the switch leads brought out to a convenient spot.

A "FAIL-PROOF" CONELRAD ALARM

The conelrad alarm shown in Fig. 5-65 uses a small BC receiver to furnish both audible and visible indications of a Conelrad Alert (the receiver may still be used for normal broadcast reception).

With the receiver tuned to a broadcast carrier and the alarm circuit in operation, a green "safe" light indicates that all is well on the broadcast band. When the broadcast carrier goes off, as it will in a Conelrad Radio Alert, the green light goes out, a red "danger" light comes on, a buzzer sounds, and the 115-volt a.c. line to the transmitter is opened up. In other words, the device *puts* you off the air! The audible and visible warnings also are given in the event of a component failure in either the control receiver or the alarm. Even the disappearance of the 115-volt supply will not go unnoticed, since in that case the green "safe" light will go out, indicating that the alarm is inoperative.

The alarm requires a minimum of 0.7 volts (negative) from the receiver's a.v.c. (automatic volume control) circuit for dependable operation. Receivers having one stage of i.f. amplification will develop at least this much a.v.c. voltage when tuned to a signal of reasonable strength. But watch out for the "superhets" that do not have an i.f. stage; they are of little value as a source of control voltage for the alarm. You can usually find out if the receiver has an i.f. stage by looking at the tube list pasted on either the chassis or the inside of the cabinet.

The circuit of the alarm is shown in section B, Fig. 5-66. Section A is a typical a.v.c.-detector-first audio stage of an a.c.-d.c. receiver, and shows how the alarm circuit is tied into a receiver.

Although a 12AV6 is shown as the detector, other tubes may be used in some receivers. However, the basic circuit will be the same or very similar.

Finding the a.v.c. line in the jumble beneath the chassis of the ordinary a.c.-d.c. receiver is not always easy. Here are a few hints:

Using section A, Fig. 5-66, as a guide, locate the detector tube socket. Trace out the leads going to the secondary of the last i.f. transformer, T_1. This transformer usually will be adjacent to the detector tube. The lower end of the secondary winding will be connected to several different resistors, one of these being the diode-load filter resistor (approximately 50K in most circuits) and another the a.v.c. filter resistor, R_1. The value of the latter resistor is ordinarily above one megohm. Trace through R_1 in the direction of the arrow (Fig. 5-66), until you locate the fairly high

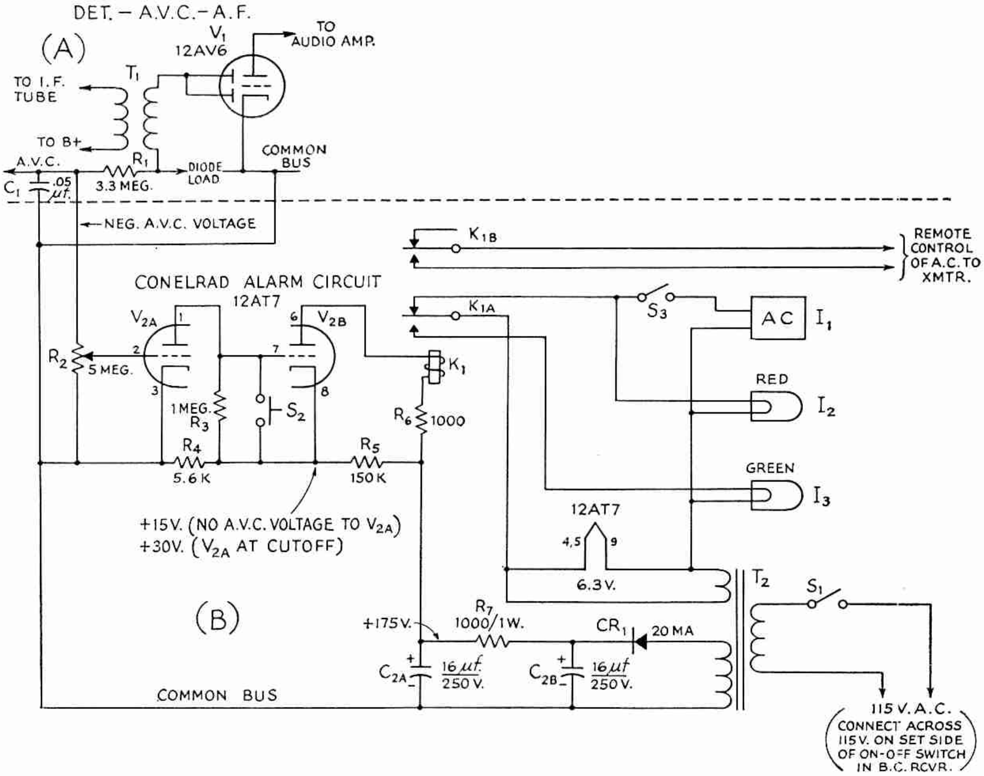


Fig. 5-66—Circuit of the Conelrad alarm (B) connected to the a.v.c. circuit (A) of a typical a.c.-d.c. broadcast receiver Resistors are ½ watt unless otherwise specified. C_1, R_1 and T_1 in section A are components in the broadcast receiver

I_1—6-volt a.c. buzzer (Edwards 725).
I_2, I_3—6-volt pilot lamp, No. 47.
K_1—D.p.d.t. sensitive relay, 5000-ohm coil, 5-amp. contacts (Potter & Brumfield GB11D).
R_2—5-megohm potentiometer.
S_1, S_2—S.p.s.t. rotary canopy switch (ICA 1257).
S_2—Momentary-contact switch (Switchcraft 101).
T_2—Replacement-type power transformer, 150 volts, 25 ma.; 6.3 volts, 0.5 amp. (Merit P-3046 or equivalent).

value (0.05 µf. or so) a.v.c. filter capacitor, C_1. Now you have the a.v.c. line clearly identified and the tap for the alarm circuit may be made.

Notice that the cathode of V_1 and the cold side of C_1 are both returned to a common bus or −B line, not directly to the chassis. Also observe that the return for the alarm circuit is made to the common bus in the receiver, not to the chassis of the set. *Do not ground this lead to the chassis or connect it to any exposed metal parts.* If there is any difficulty in locating the common bus in the vicinity of the detector stage, check back from the negative side of the power-supply filter capacitors, as this point is always attached to the common bus.

The monitor should be built in an insulated box of some kind and not in a metal case. The box can be made of plywood, or a bakelite instrument case (e.g., ICA type 8202). The bakelite case is ideal for the application, but it must be handled with care during construction, to avoid scratching, chipping, or breakage. Be especially careful when drilling large holes such as those used in mounting the pilot-lamp assemblies and switches, because a large drill tends to bind and crack the case.

Testing and Operating

The chances are pretty good that right after the receiver and the monitor have been turned on the red lamp will light and — if you haven't had the foresight to open S_3 to prevent the noise — the buzzer will sound. Tune the receiver to a broadcast station and see if the red light goes out and the green light comes on. If this happens, close S_3 and you're all set for conelrad compliance. If the "safe" light does not come on, tune around for a signal strong enough to actuate the alarm. Should the signal of greatest apparent strength fail to trigger the monitor, leave the receiver tuned to this signal and then momentarily press S_2. The alarm should now lock on "safe," provided the a.v.c. circuit delivers 0.7 volt or more to V_{2A}.

The only d.c. measurements of any consequence that need be made in checking through the alarm circuit are the output voltage of the power supply and the voltage at the cathode of V_{2B}. The proper voltages at these two points are given on the circuit diagram. If the alarm fails to respond properly, it may be advisable to check the a.v.c. voltage with a v.t.v.m.

CHAPTER 6

High-Frequency Transmitters

The principal requirements to be met in c.w. transmitters for the amateur bands between 1.8 and 30 Mc. are that the frequency must be as stable as good practice permits, the output signal must be free from modulation and that harmonics and other spurious emissions must be eliminated or reduced to the point where they do not cause interference to other stations.

The over-all design depends primarily upon the bands in which operation is desired, and the power output. A simple oscillator with satisfactory frequency stability may be used as a transmitter at the lower frequencies, as indicated in Fig. 6-1A, but the power output obtainable is small. As a general rule, the output of the oscillator is fed into one or more amplifiers to bring the power fed to the antenna up to the desired level, as shown in B.

An amplifier whose output frequency is the same as the input frequency is called a **straight amplifier.** A **buffer amplifier** is the term sometimes applied to an amplifier stage to indicate that its primary purpose is one of isolation, rather than power gain.

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. From the viewpoint of any particular stage in a transmitter, the preceding stage is its **driver.**

As a general rule, frequency multipliers should not be used to feed the antenna system directly, but should feed a straight amplifier which, in turn, feeds the antenna system, as shown in Fig. 1-C, D and E. As the diagrams indicate, it is often possible to operate more than one stage from a single power supply.

Good frequency stability is most easily obtained through the use of a **crystal-controlled oscillator,** although a different crystal is needed for each frequency desired (or multiples of that frequency). A **self-controlled oscillator** or **v.f.o.** (variable-frequency oscillator) may be tuned to any frequency with a dial in the manner of a receiver, but requires great care in design and construction if its stability is to compare with that of a crystal oscillator.

In all types of transmitter stages, screen-grid tubes have the advantage over triodes that they require less driving power. With a lower-power exciter, the problem of harmonic reduction is made easier. Most satisfactory oscillator circuits use a screen-grid tube.

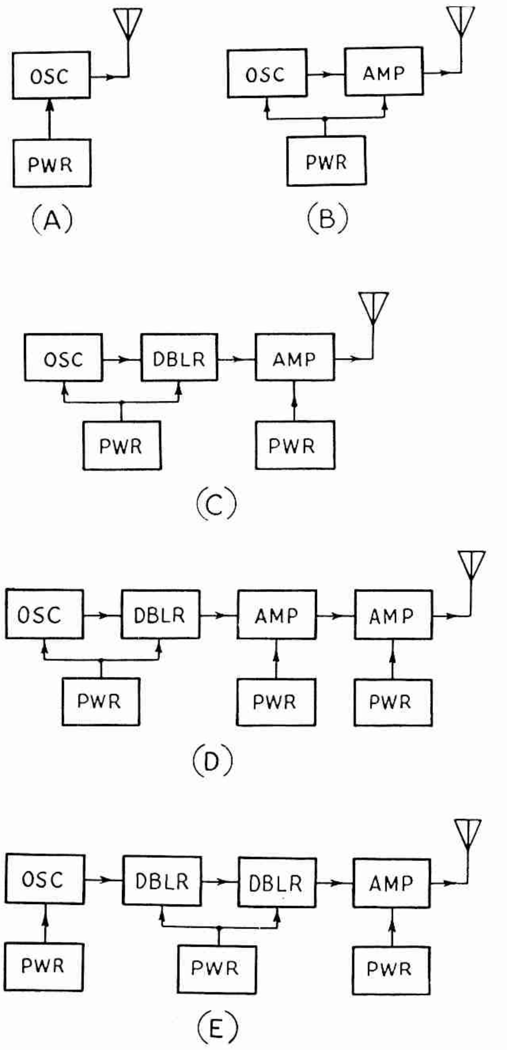


Fig. 6-1—Block diagrams showing typical combinations of oscillator and amplifiers and power-supply arrangements for transmitters. A wide selection is possible, depending upon the number of bands in which operation is desired and the power output.

6—HIGH-FREQUENCY TRANSMITTERS

Oscillators

CRYSTAL OSCILLATORS

The frequency of a crystal-controlled oscillator is held constant to a high degree of accuracy by the use of a quartz crystal. The frequency depends almost entirely on the dimensions of the crystal (essentially its thickness); other circuit values have comparatively negligible effect. However, the power obtainable 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 feedback 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.

Crystal-Oscillator Circuits

The simplest crystal-oscillator circuit is shown in Fig. 6-2A. An equivalent is shown at B. It is a Colpitts circuit (see chapter on vacuum-tube principles) with the tube tapped across part of the tuned circuit. The crystal has been replaced by its equivalent — a series-tuned circuit L_1C_4. (See chapter on electrical laws and circuits.) C_5 and C_6 are the tube grid-cathode and plate-cathode capacitances, respectively. In best practical form, C_5 or C_6, or both, would be augmented by external capacitors from grid to cathode and plate to cathode so that feedback could be adjusted properly.

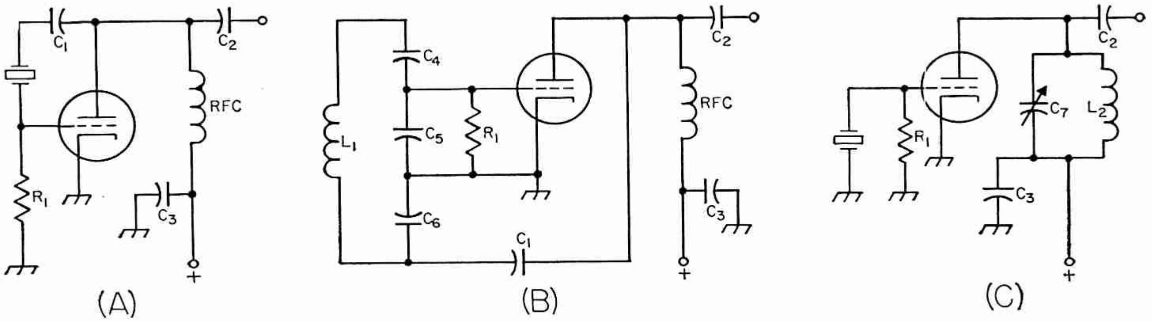


Fig. 6-2—Simple crystal-oscillator circuits. A—Pierce. B—Equivalent of circuit A. C—Simple triode oscillator. C_1 is a plate blocking capacitor, C_2 an output coupling capacitor, and C_3 a plate bypass. L_1, C_4, C_5 and C_6 are discussed in the text. C_7 and L_2 should tune to the crystal fundamental frequency. R_1 is the grid leak.

The circuit shown in Fig. 6-2C is the equivalent of the tuned-grid tuned-plate circuit discussed in the chapter on vacuum-tube principles, the crystal replacing the tuned grid circuit

The most commonly used crystal-oscillator circuits are based on one or the other of these two simple types, and are shown in Fig. 6-3. Although these circuits are somewhat more complicated, they combine the functions of oscillator and amplifier or frequency multiplier in a single tube. In all of these circuits, the screen of a tetrode or pentode is used as the plate in a triode oscillator. Power output is taken from a separate tuned tank circuit in the actual plate circuit. Although the oscillator itself is not entirely independent of adjustments made in the plate tank circuit when the latter is tuned near the fundamental frequency of the crystal, the effects can be satisfactorily minimized by proper choice of the oscillator tube.

The circuit of Fig. 6-3A is known as the Tri-tet. The oscillator circuit is that of Fig. 6-2C. Excitation is controlled by adjustment of the tank L_1C_1, which should have a low L/C ratio, and be tuned considerably to the high-frequency side of the crystal frequency (approximately 5 Mc. for a 3.5-Mc. crystal) to prevent over-excitation and high crystal current. Once the proper adjustment for average crystals has been found, C_1 may be replaced with a fixed capacitor of equal value.

The oscillator circuit of Fig. 3-B is that of Fig. 6-2A. Excitation is controlled by C_9.

The oscillator of the grid-plate circuit of Fig. 6-3C is the same as that of Fig. 6-3B, except that the ground point has been moved from the cathode to the plate of the oscillator (in other words, to the screen of the tube). Excitation is adjusted by proper proportioning of C_6 and C_7.

When most types of tubes are used in the circuits of Fig. 6-3, oscillation will stop when the output plate circuit is tuned to the crystal frequency, and it is necessary to operate with the plate tank circuit critically detuned for maximum output with stability. However, when the 6AG7, 5763, or the lower-power 6AH6 is used with proper adjustment of excitation, it is possible to tune to the crystal frequency without stopping oscillation. The plate tuning characteristic should then be similar to Fig. 6-4. These tubes also operate with less crystal current than most other types for a given power output, and less frequency change occurs when the plate circuit is tuned through the crystal frequency (less than 25 cycles at 3.5 Mc.).

Crystal current may be estimated by observing the relative brilliance of a 60-ma. dial lamp connected in series with the crystal. Current should be held to the minimum for satisfactory output by careful adjustment of excitation. With the operating voltages shown, satisfactory output should be obtained with crystal currents of 40 ma. or less.

In these circuits, output may be obtained at multiples of the crystal frequency by tuning the plate tank circuit to the desired harmonic, the output dropping off, of course, at the higher harmonics. Especially for harmonic operation, a low-*C* plate tank circuit is desirable.

For best performance with a 6AG7 or 5763, the values given under Fig. 6-3 should be followed closely. (For a discussion of values for other tubes, see *QST* for March, 1950, page 28.)

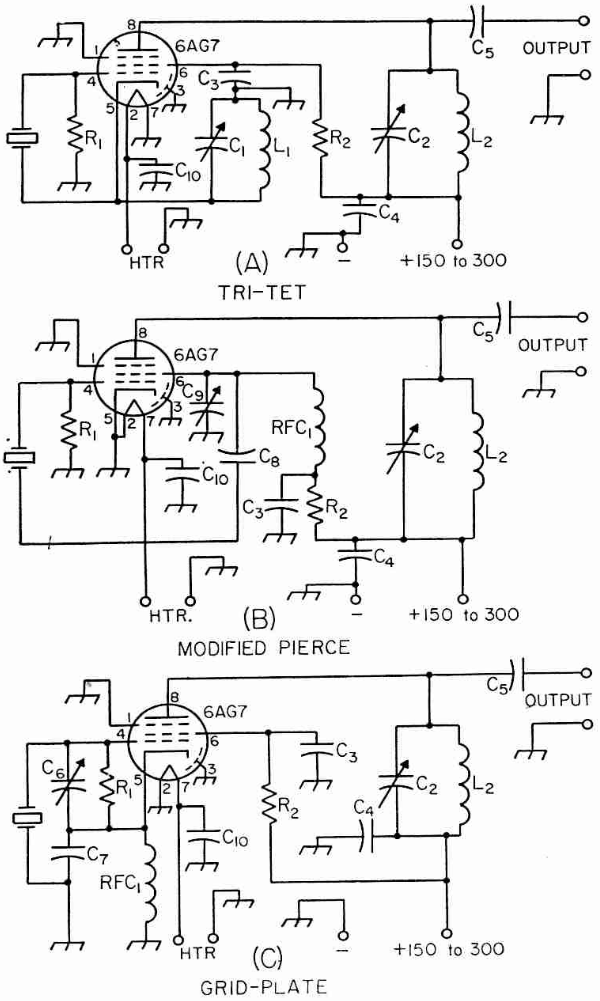


Fig. 6-3—Commonly used crystal-controlled oscillator circuits. Values are those recommended for a 6AG7 or 5763 tube. (See reference in text for other tubes.)

C_1—Feedback-control capacitor—3.5-Mc. crystals—approx. 220-μμf. mica—7-Mc. crystals—approx. 150-μμf. mica.

C_2—Output tank capacitor—100-μμf. variable for single-band tank; 250-μμf. variable for two-band tank.

C_3—Screen bypass—0.001-μf. disk ceramic.

C_4—Plate bypass—0.001-μf. disk ceramic.

C_5—Output coupling capacitor—50 to 100 μμf.

C_6—Excitation-control capacitor—30-μμf. trimmer.

C_7—Excitation capacitor—220-μμf. mica for 6AG7; 100-μμf. for 5763.

C_8—D.c. blocking capacitor—0.001-μf. mica.

C_9—Excitation-control capacitor—220-μμf. mica.

C_{10}—Heater bypass—0.001-μf. disk ceramic.

R_1—Grid leak—0.1 megohm, ½ watt.

R_2—Screen resistor—47,000 ohms, 1 watt.

L_1—Excitation-control inductance—3.5-Mc. crystals—approx. 4 μh.; 7-Mc. crystals—approx. 2 μh.

L_2—Output-circuit coil—single band:—3.5 Mc.—17 μh.; 7 Mc.—8 μh.; 14 Mc.—2.5 μh.; 28 Mc.—1 μh. Two-band operation: 3.5 & 7 Mc.—7.5 μh.; 7 & 14 Mc.—2.5 μh.

RFC$_1$—2.5-mh. 50-ma. r.f. choke.

VARIABLE-FREQUENCY OSCILLATORS

The frequency of a v.f.o. depends entirely on the values of inductance and capacitance in the circuit. Therefore, it is necessary to take careful steps to minimize changes in these values not under the control of the operator. As examples, even the minute changes of dimensions with temperature, particularly those of the coil, may result in a slow but noticeable change in frequency called **drift.** The effective input capacitance of the oscillator tube, which must be connected across the circuit, changes with variations in electrode voltages. This, in turn, causes a change in the frequency of the oscillator. To make use of the power from the oscillator, a load, usually in the form of an amplifier, must be coupled to the oscillator, and variations in the load may reflect on the frequency. Very slight mechanical movement of components may result in a shift in frequency, and vibration can cause modulation.

V.F.O. Circuits

Fig. 6-5 shows the most commonly used circuits. They are all designed to minimize the effects mentioned above. All are similar to the crystal oscillators of Fig. 6-3 in that the screen of a tetrode or pentode is used as the oscillator plate. The oscillating circuits in Figs. 6-5A and B are the Hartley type; those in C and D are Colpitts circuits. (See chapter on vacuum-tube principles.) In the circuits of A and C, all of the above-mentioned effects, except changes in inductance, are minimized by the use of a high-*Q* tank circuit obtained through the use of large tank capacitances. Any uncontrolled changes in capacitance thus become a very small percentage of the total circuit capacitance.

In the series-tuned Colpitts circuit of Fig. 6-5D (sometimes called the Clapp circuit), a high-*Q* circuit is obtained in a different manner. The tube is tapped across only a small portion of the oscillating tank circuit, resulting in very loose coupling between tube and circuit. The taps are provided by a series of three capacitors across the coil. In addition, the tube capacitances are shunted by large capacitors, so the effects of the tube — changes in electrode voltages and loading — are still further reduced. In contrast

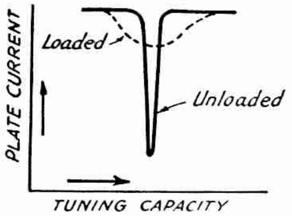


Fig. 6-4 — Plate tuning characteristic of circuits of Fig. 6-3 with preferred types (see text). The plate-current dip at resonance broadens and is less pronounced when the circuit is loaded.

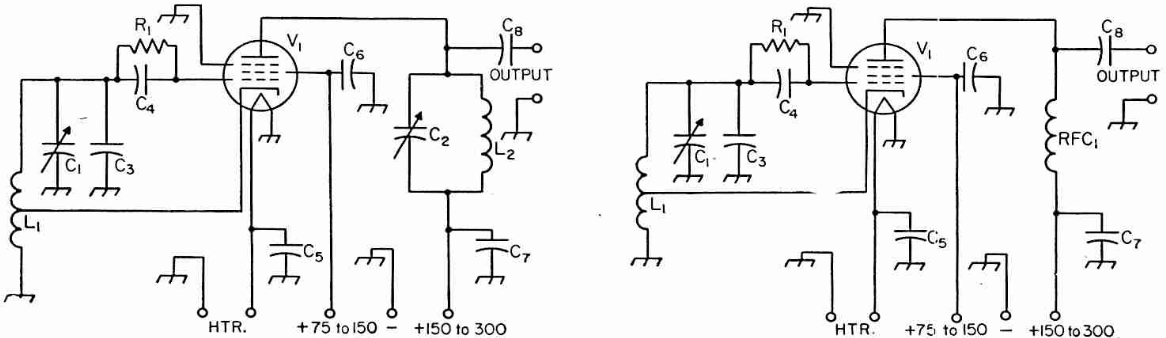
to the preceding circuits, the resulting tank circuit has a high L/C ratio and therefore the tank current is much lower than in the circuits using high-C tanks. As a result, it will usually be found that, other things being equal, drift will be less with the low-C circuit.

For best stability, the ratio of C_{13} or C_{14} (which are usually equal) to $C_{11} + C_{12}$ should be as high as possible without stopping oscillation. The permissible ratio will be higher the higher the Q of the coil and the mutual conductance of the tube. If the circuit does not oscillate over the desired range, a coil of higher Q must be used or the capacitance of C_{13} and C_{14} reduced.

Load Isolation

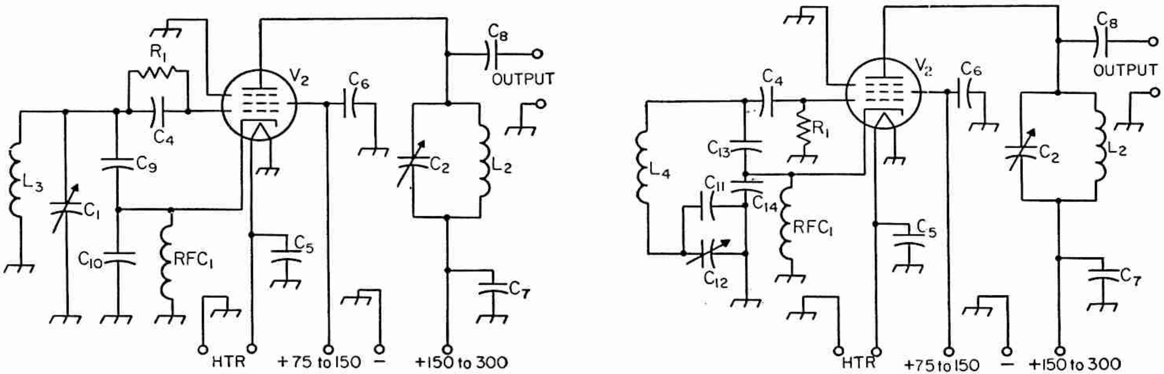
In spite of the precautions already discussed, the tuning of the output plate circuit will cause a noticeable change in frequency, particularly in the region around resonance. This effect can be reduced considerably by designing the oscillator for half the desired frequency and doubling frequency in the output circuit.

It is desirable, although not a strict necessity if detuning is recognized and taken into account, to approach as closely as possible the condition where the adjustment of tuning controls in the transmitter, beyond the v.f.o. frequency control, will have negligible effect on the frequency. This can be done by substituting a fixed-tuned circuit in the output of the oscillator, and adding isolating stages whose tuning is fixed between the oscillator and the first tunable amplifier stage in the transmitter. Fig. 6-6 shows such an arrangement that gives good isolation. In the first stage, a 6C4 is connected as a cathode follower. This



(A) HARTLEY

(B) HARTLEY NONRESONANT OUTPUT



(C) COLPITTS

(D) SERIES-TUNED COLPITTS

Fig. 6-5—V.f.o. circuits. Approximate values for 3.5 Mc. are given below. For 1.75 Mc., all tank-circuit values of capacitance and inductance, all tuning capacitances and C_{13} and C_{14} should be doubled; for 7 Mc., they should be cut in half.

C_1—Oscillator bandspread tuning capacitor—150-μμf. variable.

C_2—Output-circuit tank capacitor—100-μμf.

C_3—Oscillator tank capacitor—500-μμf. zero-temperature-coefficient mica.

C_4—Grid coupling capacitor—100-μμf. zero-temperature-coefficient mica.

C_5—Heater bypass—0.001-μf. disk ceramic.

C_6—Screen bypass—0.001-μf. disk ceramic.

C_7—Plate bypass—0.001-μf. disk ceramic.

C_8—Output coupling capacitor—50 to 100-μμf. mica.

C_9—Oscillator tank capacitor—680-μμf. zero-temperature-coefficient mica.

C_{10}—Oscillator tank capacitor—0.0022-μf. zero-temperature-coefficient mica.

C_{11}—Oscillator bandspread padder—50-μμf. variable air.

C_{12}—Oscillator bandspread tuning capacitor—25-μμf. variable.

C_{13}, C_{14}—Tube-coupling capacitor—0.001-μf. zero-temperature-coefficient mica.

R_1—47,000 ohms, ½ watt.

L_1—Oscillator tank coil—4.3 μh., tapped about one-third-way from grounded end.

L_2—Output-circuit tank coil—22 μh.

L_3—Oscillator tank coil—4.3 μh.

L_4—Oscillator tank coil—33 μh. (B & W JEL-80).

RFC$_1$—2.5-mh. 50-ma. r.f. choke.

V_1—6AG7, 5763 or 6AH6 preferred; other types usable.

V_2—6AG7, 5763 or 6AH6 required for feedback capacitances shown.

drives a 5763 buffer amplifier whose input circuit is fixed-tuned to the approximate band of the v.f.o. output. For best isolation, it is important that the 6C4 does not draw grid current. The output of the v.f.o., or the cathode resistor of the 6C4 should be adjusted until the voltage across the cathode resistor of the 6C4 (as measured with a high-resistance d.c. voltmeter with an r.f. choke in the positive lead) is the same with or without excitation from the v.f.o. L_1 should be adjusted for most constant output from the 5763 over the band.

Chirp

In all of the circuits shown there will be some change of frequency with changes in screen and plate voltages, and the use of regulated voltages for both usually is necessary. One of the most serious results of voltage instability occurs if the oscillator is keyed, as it often is for break-in operation. Although voltage regulation will supply a steady voltage from the power supply and therefore is still desirable, it cannot alter the fact that the voltage on the tube must rise from zero when the key is open, to full voltage when the key is closed, and must fall back again to zero when the key is opened. The result is a chirp each time the key is opened or closed, unless the time constant in the keying circuit is reduced to the point where the chirp takes place so rapidly that the receiving operator's ear cannot detect it. Unfortunately, as explained in the chapter on keying, a certain minimum time constant is necessary if key clicks are to be minimized. Therefore it is evident that the measures necessary for the reduction of chirp and clicks are in opposition, and a compromise is necessary. For best keying characteristics, the oscillator should be allowed to run continuously while a subsequent amplifier is keyed. However, a keyed amplifier represents a widely variable load and unless sufficient isolation is provided between the oscillator and the keyed amplifier, the keying characteristics may be little better than when the oscillator itself is keyed. (See keying chapter for other methods of break-in keying.)

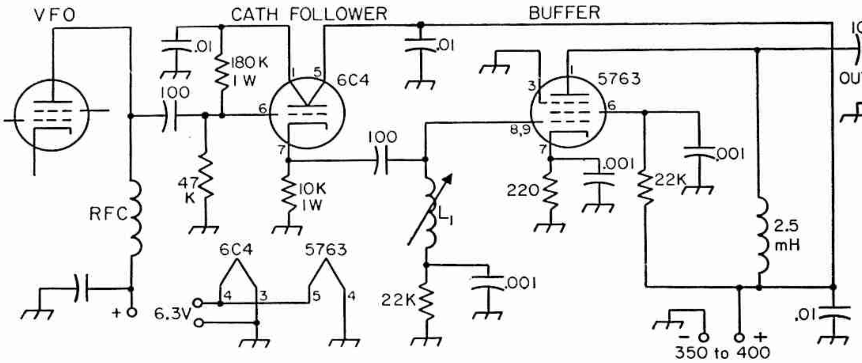


Fig. 6-6—Circuit of an isolating amplifier for use between v.f.o. and first tunable stage. All capacitances below 0.001 μf. are in $\mu\mu$f. All resistors are ½ watt. L_1, for the 3.5-Mc. band, consists of 93 turns No. 36 enam., 17/32 inch long, ½-inch diameter, close-wound on National XR-50 iron-slug form. Inductance 69 to 134 μh. All capacitors are disk ceramic.

Frequency Drift

Frequency drift is further reduced most easily by limiting the power input as much as possible and by mounting the components of the tuned circuit in a separate shielded compartment, so that they will be isolated from the direct heat from tubes and resistors. The shielding also will eliminate changes in frequency caused by movement of nearby objects, such as the operator's hand when tuning the v.f.o. The circuit of Fig. 6-5D lends itself well to this arrangement, since relatively long leads between the tube and the tank circuit have negligible effect on frequency because of the large shunting capacitances. The grid, cathode and ground leads to the tube can be bunched in a cable up to several feet long.

Variable capacitors should have ceramic insulation, good bearing contacts and should preferably be of the double-bearing type, and fixed capacitors should have zero temperature coefficient. The tube socket also should have ceramic insulation and special attention should be paid to the selection of the coil in the oscillating section.

Oscillator Coils

The Q of the tank coil used in the oscillating portion of any of the circuits under discussion should be as high as circumstances (usually space) permit, since the losses, and therefore the heating, will be less. With recommended care in regard to other factors mentioned previously, most of the drift will originate in the coil. The coil should be well spaced from shielding and other large metal surfaces, and be of a type that radiates heat well, such as a commercial air-wound type, or should be wound tightly on a threaded ceramic form so that the dimensions will not change readily with temperature. The wire with which the coil is wound should be as large as practicable, especially in the high-C circuits.

Mechanical Vibration

To eliminate mechanical vibration, components should be mounted securely. Particularly in the circuit of Fig. 6-5D, the capacitor should preferably have small, thick plates and the coil braced, if necessary, to prevent the slightest mechanical movement. Wire connections between tank-circuit components should be as short as possible and flexible wire will have less tendency to vibrate than solid wire. It is advisable to cushion the entire oscillator unit by mounting on sponge rubber or other shock mounting.

Tuning Characteristic

If the circuit is oscillating, touching the grid of the tube or any part of the circuit connected to it will show a change in plate current. In tuning the plate output circuit without load, the plate current will be relatively high until it is tuned near resonance where the plate current will dip to a low value, as illustrated in Fig. 6-4. When the output circuit is loaded, the dip should still be found, but broader and much less pronounced as indicated by the dashed line. The circuit should not be loaded beyond the point where the dip is still recognizable.

Checking V.F.O. Stability

A v.f.o. 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 v.f.o. will show signals of good quality and stability when it is running free and not connected to a load. A well-isolated monitor is a necessity. Perhaps the most convenient, as well as one of the most satisfactory, well-shielded monitoring arrangements is a receiver combined with a crystal oscillator, as shown in Fig. 6-7. (See "Crystal Oscillators," 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 v.f.o. 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 crystals have a sufficiently low temperature coefficient to give a 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 a harmonic.

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.

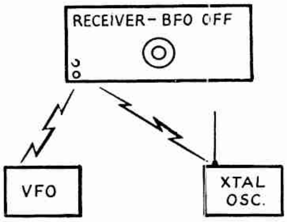


Fig. 6-7—Setup for checking v.f.o. stability. The receiver should be tuned preferably to a harmonic of the v.f.o. frequency. The crystal oscillator may operate somewhere in the band in which the v.f.o. is operating. The receiver b.f.o. should be turned off.

R.F. Power-Amplifier Tanks and Coupling

R.f. power amplifiers used in amateur transmitters usually are operated under Class C conditions (see chapter on vacuum-tube fundamentals). Fig. 6-10 shows a screen-grid tube with the required tuned tank in its plate circuit. Equivalent cathode connections for a filament-type tube are shown in Fig. 6-8 It is assumed that the tube is being properly driven and that the various electrode voltages are appropriate for Class C operation.

● PLATE TANK Q

The main objective, of course, is to deliver as much fundamental power as possible into a load, *R*, without exceeding the tube ratings. The load resistance *R* may be in the form of a transmission line to an antenna, or the grid circuit of another amplifier. A further objective is to minimize the harmonic energy (always generated by a Class C amplifier) fed into the load circuit. In attaining these objectives, the *Q* of the tank circuit is of importance. When a load is coupled inductively, as in Fig. 6-10, the *Q* of the tank circuit will have an effect on the coefficient of coupling necessary for proper loading of the amplifier. In respect to all of these factors, a tank *Q* of 10 to 20 is usually considered optimum. A much lower *Q* will result in less efficient operation of the amplifier tube, greater harmonic output, and greater difficulty in coupling inductively to a load. A much higher *Q* will result in higher tank current with increased loss in the tank coil.

The *Q* is determined (see chapter on electrical laws and circuits) by the *L/C* ratio and the load resistance at which the tube is operated. The tube load resistance is related, in approximation, to

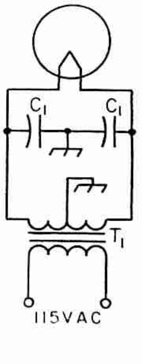


Fig. 6-8—Filament center-tap connections to be substituted in place of cathode connections shown in diagrams when filament-type tubes are substituted. T_1 is the filament transformer. Filament bypasses, C_1, should be 0.001-μf. disk ceramic capacitors. If a self-biasing (cathode) resistor is used, it should be placed between the center tap and ground.

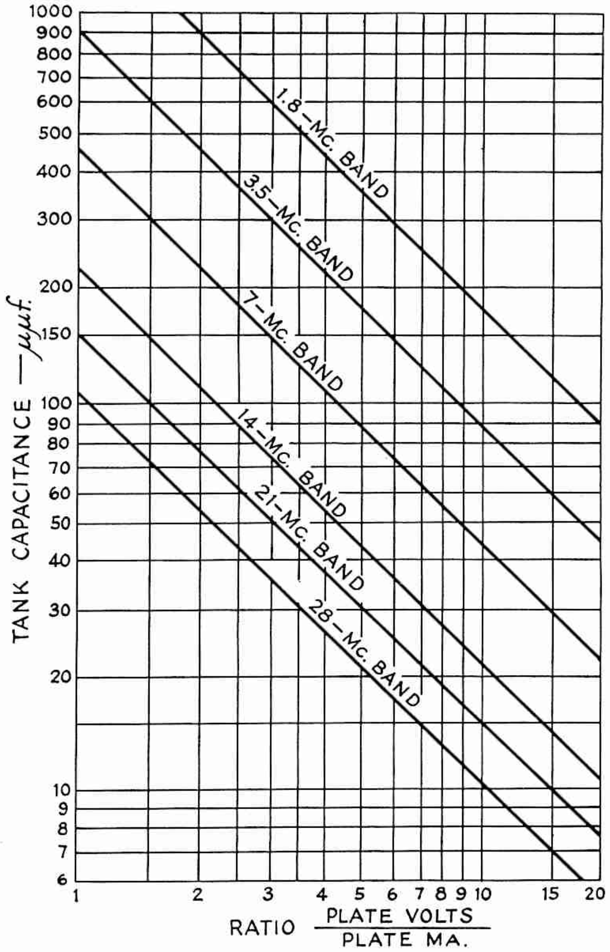


Fig. 6-9—Chart showing plate tank capacitance required for a Q of 10. Divide the tube plate voltage by the plate current in milliamperes. Select the vertical line corresponding to the answer obtained. Follow this vertical line to the diagonal line for the band in question, and thence horizontally to the left to read the capacitance. For a given ratio of plate-voltage/plate current, doubling the capacitance shown doubles the Q etc. When a split-stator capacitor is used in a balanced circuit, the capacitance of each section may be one half of the value given by the chart.

the ratio of the d.c. plate voltage to d.c. plate current at which the tube is operated.

The amount of *C* that will give a *Q* of 10 for various ratios is shown in Fig. 6-9. For a given plate-voltate/plate-current ratio, the *Q* will vary directly as the tank capacitance, twice the capacitance doubles the *Q* etc. For the same *Q*, the capacitance of *each section* of a split-stator capacitor in a balanced circuit should be half the value shown.

These values of capacitance include the output capacitance of the amplifier tube, the input capacitance of a following amplifier tube if it is coupled capacitively, and all other stray capacitances. At the higher plate-voltage/plate-current ratios, the chart may show values of capacitance, for the higher frequencies, smaller than those attainable in practice. In such a case, a tank *Q* higher than 10 is unavoidable.

In low-power exciter stages, where capacitive coupling is used, very low-*Q* circuits, tuned only by the tube and stray circuit capacitances are sometimes used for the purpose of "broadbanding" to avoid the necessity for retuning a stage across a band. Higher-order harmonics generated in such a stage can usually be attentuated in the tank circuit of the final amplifier.

● INDUCTIVE-LINK COUPLING

Coupling to Flat Coaxial Lines

When the load *R* in Fig. 6-10 is located for convenience at some distance from the amplifier, or when maximum harmonic reduction is desired, it is advisable to feed the power to the load through a low-impedance coaxial cable. The shielded construction of the cable prevents radiation and makes it possible to install the line in any convenient manner without danger of unwanted coupling to other circuits.

If the line is more than a small fraction of a wavelength long, the load resistance at its output end should be adjusted, by a matching circuit if necessary, to match the impedance of the cable. This reduces losses in the cable and makes the coupling adjustments at the transmitter independent of the cable length. Matching circuits for use between the cable and another transmission line are discussed in the chapter on transmission lines, while the matching adjustments when the load is the grid circuit of a following amplifier are described elsewhere in this chapter.

Assuming that the cable is properly terminated, proper loading of the amplifier will be assured, using the circuit of Fig. 6-11C, if

1) The plate tank circuit has reasonably high value of *Q*. A value of 10 is usually sufficient.

2) The inductance of the pick-up or link coil is 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 charac-

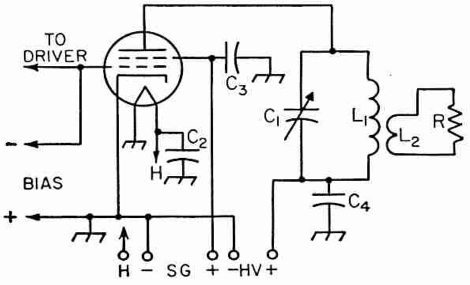


Fig. 6-10—Inductive-link output coupling circuits.

C_1—Plate tank capacitor—see text and Fig. 6-9 for capacitance, Fig. 6-33 for voltage rating.

C_2—Heater bypass—0.001-μf. disk ceramic.

C_3—Screen bypass—voltage rating depends on method of screen supply. See paragraphs on screen considerations. Voltage rating same as plate voltage will be safe under any condition.

C_4—Plate bypass—0.001-μf. disk ceramic or mica. Voltage rating same as C_1, plus safety factor.

L_1—To resonate at operating frequency with C_1. See LC chart and inductance formula in electrical-laws chapter, or use ARRL *Lightning Calculator*.

L_2—Reactance equal to line impedance. See reactance chart and inductance formula in electrical-laws section, or use ARRL *Lightning Calculator*.

R—Representing load.

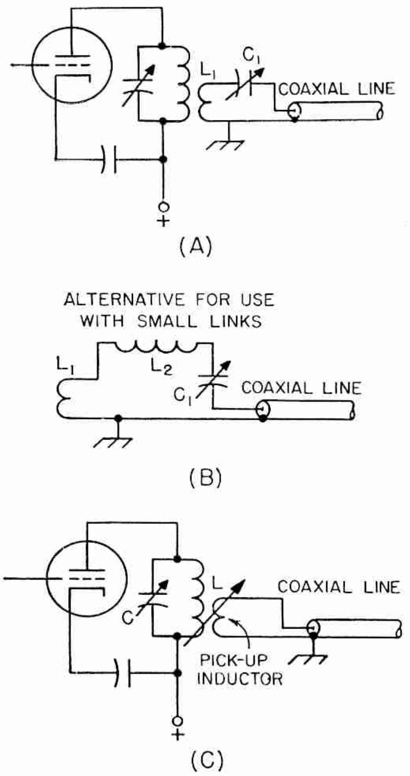


Fig. 6-11—With flat transmission lines power transfer is obtained with looser coupling if the line input is tuned to resonance. C_1 and L_1 should resonate at the operating frequency. See table for maximum usable value of C_1. If circuit does not resonate with maximum C_1 or less, inductance of L_1 must be increased, or added in series at L_2.

teristic impedance, Z_0, of the line.

3) It is possible to make the coupling between the tank and pick-up coils very tight.

The second in this list is often hard to meet. Few manufactured link coils have adequate inductance even for coupling to a 50-ohm line at low frequencies.

If the line is operating with a low s.w.r., the system shown in Fig. 6-11C 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 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.

Capacitance in μμf. Required for Coupling to Flat Coaxial Lines with Tuned Coupling Circuit

Frequency Band Mc.	*Characteristic Impedance of Line*	
	52 ohms[1]	*75 ohms*[1]
1.8	900	600
3.5	450	300
7	230	150
14	115	75
28	60	40

[1] Capacitance values are maximum usable.

Note: Inductance in circuit must be adjusted to resonate at operating frequency.

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 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. With coaxial cable, a circuit of reasonable Q can be obtained with practicable values of inductance and capacitance connected in series with the line's input terminals. Suitable circuits are given in Fig. 6-11 at A and B. 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 line impedances of 52 and 75 ohms are given in the accompanying table. These are the *maximum* values that should be used. 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 may be connected in series as shown in Fig. 6-11B.

Characteristics

In practice, the amount of inductance in the circuit should be chosen so that, with somewhat loose coupling between L_1 and the amplifier tank coil, the amplifier plate current will increase when the variable capacitor, 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, without changing the setting of C_1. 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 line actually will be flat over such a range, so some readjustment of C_1 may be needed to compensate for changes in the input impedance of the line. 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. This can be done by increasing the L/C ratio.

● PI-SECTION OUTPUT TANK

A pi-section tank circuit may also be used in coupling to an antenna or transmission line, as shown in Fig. 6-12. The values of capacitance for C_1 and C_2, and inductance for L_1 for any values of tube load resistance and output load resistance may be calculated from the formulas in the chapter on electrical laws.

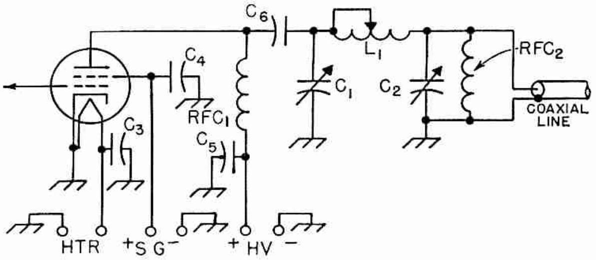


Fig. 6-12—Pi-section output tank circuit.

C_1—Input capacitor. See text or Fig. 6-13 for reactance. Voltage rating should be equal to d.c. plate voltage for c.w.; double this value for plate modulation.
C_2—Output capacitor. See text or Fig. 6-15 for reactance. See text for voltage rating.
C_3—Heater bypass—0.001-μf. disk ceramic.
C_4—Screen bypass. See Fig. 6-10.
C_5—Plate bypass. See Fig. 6-10.
C_6—Plate blocking capacitor—0.001-μf. disk ceramic or mica. Voltage rating same as C_1.
L_1—See text or Fig. 6-14 for reactance.
RFC_1—See later paragraph on r.f. chokes.
RFC_2—2.5-mh. receiving type (essential to reduce peak voltage across both input and output capacitors).

Values of reactance for C_1, L_1 and C_2 may be taken directly from the charts of Figs. 6-13, 6-14 and 6-15 if the output load resistance is 52 or 72 ohms. It should be borne in mind that these values apply only where the output load is resistive, i.e., where the antenna and line have been matched. The tube load resistance R_1 in ohms is determined by dividing the plate voltage by twice the d.c. plate current in decimal parts of an ampere.

Output-Capacitor Ratings

The voltage rating of the output capacitor will depend upon the s.w.r. If the load is resistive, receiving-type air capacitors should be adequate for amplifier input powers up to 1 kw. with plate modulation when feeding 52- or 72-ohm loads. In obtaining the larger capacitances required for the lower frequencies, it is common practice to switch fixed capacitors in parallel with the variable air capacitor. While the voltage rating of a mica or ceramic capacitor may not be exceeded in a particular case, capacitors of these types are limited in current-carrying capacity. Postage-stamp silver-mica capacitors should be adequate for amplifier inputs over the range from about 70 watts at 28 Mc. to 400 watts at 14 Mc. and lower. The larger mica capacitors (CM-45 case) having voltage ratings of 1200 and 2500 volts are usually satisfactory for inputs varying from about 350 watts at 28 Mc. to 1 kw. at 14 Mc. and lower. Because of these current limitations, particularly at the higher frequencies, it is ad-

PI-NETWORK DESIGN CHARTS FOR FEEDING 52- OR 72-OHM COAXIAL TRANSMISSION LINES

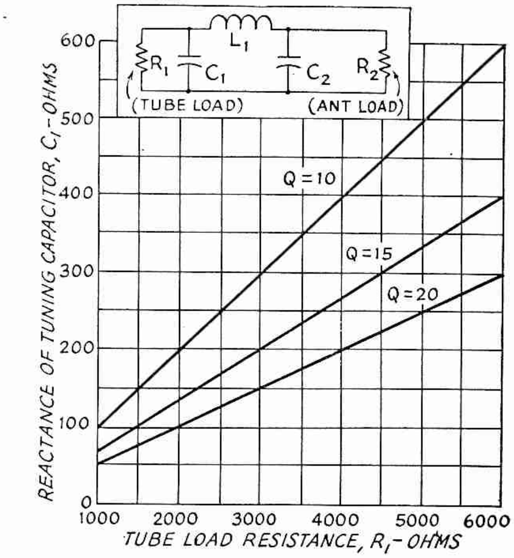


Fig. 6-13—Reactance of input capacitor, C_1, as a function of tube load resistance, R_1, for pi networks. R_1 equals plate voltage divided by twice plate current (amperes).

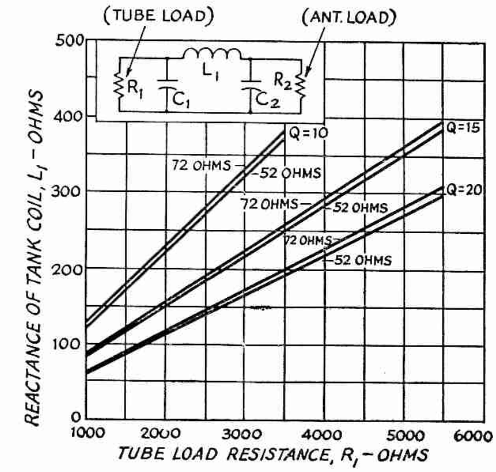


Fig. 6-14—Reactance of tank coil, L_1, as a function of load resistance. R_1, for pi networks.

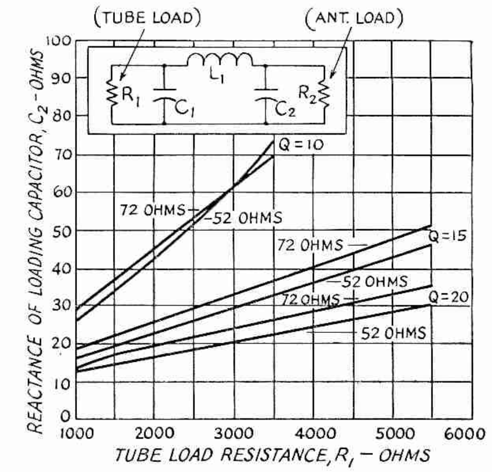


Fig. 6-15—Reactance of loading capacitor, C_2, as a function of tube load resistance, R_1, for pi networks.

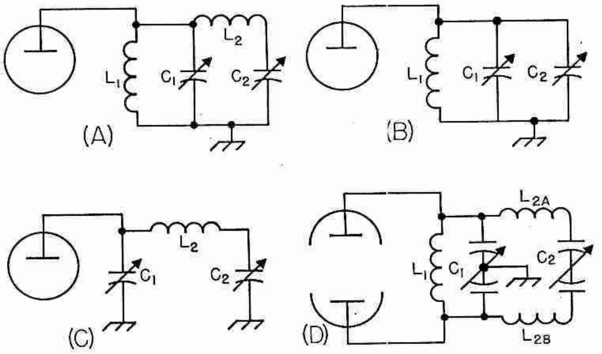


Fig. 6-16—Multiband tuner circuits. In the unbalanced circuit of A, C_1 and C_2 are sections of a single split-stator capacitor. In the balanced circuit of D, the two split-stator capacitors are ganged to a single control with an insulated shaft coupling between the two. In D, the two sections of L_2 are wound on the same form, with the inner ends connected to C_2. In A, each section of the capacitor should have a voltage rating the same as Fig. 6-33A. In D, C_1 should have a rating the same as Fig. 6-33H (or Fig. 6-33E if the feed system corresponds). C_2 may have the rating of Fig. 6-33E so long as the rotor is not grounded or bypassed to ground.

visable to use as large an air capacitor as practicable, using the micas only at the lower frequencies. Broadcast-receiver replacement-type capacitors can be obtained very reasonably. They are available in triple units totaling about 1100 $\mu\mu$f., or dual units totaling about 900 $\mu\mu$f. Their insulation should be sufficient for inputs of 500 watts or more. Air capacitors have the additional advantage that they are seldom permanently damaged by a voltage break-down.

Neutralizing with Pi Network

Screen-grid amplifiers using a pi-network output circuit may be neutralized by the system shown in Figs. 6-23B and C.

● MULTIBAND TANK CIRCUITS

Multiband tank circuits provide a convenient means of covering several bands without the need for changing coils. Tuners of this type consist essentially of two tank circuits, tuned simultaneously with a single control. In a tuner designed to cover 80 through 10 meters, each circuit has a sufficiently large capacitance variation to assure an approximately 2-to-1 frequency range. Thus, one circuit is designed so that it covers 3.5 through 7.3 Mc., while the other covers 14 through 29.7 Mc.

A single-ended, or unbalanced, circuit of this type is shown in Fig. 6-16A. In principle, the reactance of the high-frequency coil, L_2, is small enough at the lower frequencies so that it can be largely neglected, and C_1 and C_2 are in parallel across L_1. Then the circuit for low frequencies becomes that shown in Fig. 6-16B.

At the high frequencies, the reactance of L_1 is high, so that it may be considered simply as a choke shunting C_1. The high-frequency circuit is essentially that of Fig. 6-16C, L_2 being tuned by C_1 and C_2 in series.

In practice, the effect of one circuit on the other cannot be neglected entirely. L_2 tends to increase the effective capacitance of C_2, while L_1 tends to decrease the effective capacitance of C_1. This effect, however, is relatively small. Each circuit must cover somewhat more than a 2-to-1 frequency range to permit staggering the two ranges sufficiently to avoid simultaneous responses to a frequency in the low-frequency range, and one of its harmonics lying in the range of the high-frequency circuit.

In any circuit covering a frequency range as great as 2 to 1 by capacitance alone, the circuit Q must vary rather widely. If the circuit is designed for a Q of 12 at 80, the Q will be 6 at 40, 24 at 20, 18 at 15, and 12 at 10 meters. The increase in tank current as a result of the increase in Q toward the low-frequency end of the high-frequency range may make it necessary to design the high-frequency coil with care to minimize loss in this portion of the tuning range. It is generally found desirable to provide separate output coupling coils for each circuit.

Fig. 6-16D shows a similar tank for balanced circuits. The same principles apply.

Series or parallel feed may be used with either balanced or unbalanced circuits. In the balanced circuit of Fig. 6-16D, the series feed point would be at the center of L_1, with an r.f. choke in series.

(For further discussion see *QST*, July, 1954.)

R.F. Amplifier-Tube Operating Conditions

In addition to proper tank and output-coupling circuits discussed in the preceding sections, an r.f. amplifier must be provided with suitable electrode voltages and an r.f. driving or excitation voltage (see vacuum-tube chapter).

All r. f. amplifier tubes require a voltage to operate the filament or heater (a.c. is usually permissible), and a positive d.c. voltage between the plate and filament or cathode (plate voltage). Most tubes also require a negative d.c. voltage (biasing voltage) between control grid (Grid No. 1) and filament or cathode. Screen-grid tubes require in addition a positive voltage (screen voltage or Grid No. 2 voltage) between screen and filament or cathode.

Biasing and plate voltages may be fed to the tube either in series with or in parallel with the associated r.f. tank circuit as discussed in the chapter on electrical laws and circuits.

It is important to remember that true plate, screen or biasing voltage is the voltage between the particular electrode and filament or cathode. Only when the cathode is directly grounded to the chassis may the electrode-to-chassis voltage be taken as the true voltage.

The required r.f. driving voltage is applied between grid and cathode.

Power Input and Plate Dissipation

Plate power input is the d.c. power input to the plate circuit (d.c. plate voltage × d.c. plate current. Screen power input likewise is the d.c. screen voltage × the d.c. screen current.

Plate dissipation is the difference between the r.f. power delivered by the tube to its loaded plate tank circuit and the d.c. plate power input. The screen, on the other hand, does not deliver any output power, and therefore its dissipation is the same as the screen power input.

TRANSMITTING-TUBE RATINGS

Tube manufacturers specify the maximum values that should be applied to the tubes they produce. They also publish sets of typical operating values that should result in good efficiency and normal tube life.

Maximum values for all of the most popular transmitting tubes will be found in the tables of transmitting tubes in the last chapter. Also included are as many sets of typical operating values as space permits. However, it is recommended that the amateur secure a transmitting-tube manual from the manufacturer of the tube or tubes he plans to use.

CCS and ICAS Ratings

The same transmitting tube may have different ratings depending upon the manner in which the tube is to be operated, and the service in which it is to be used. These different ratings are based primarily upon the heat that the tube can safely dissipate. Some types of operation, such as with grid or screen modulation, are less efficient than others, meaning that the tube must dissipate more heat. Other types of operation, such as c.w. or single-sideband phone are intermittent in nature, resulting in less average heating than in other modes where there is a continuous power input to the tube during transmissions. There are also different ratings for tubes used in transmitters that are in almost constant use (CCS — Continuous Commercial Service), and for tubes that are to be used in transmitters that average only a few hours of daily operation (ICAS — Intermittent Commercial and Amateur Service). The latter are the ratings used by amateurs who wish to obtain maximum output with reasonable tube life.

Maximum Ratings

Maximum ratings, where they differ from the values given under typical operating values, are not normally of significance to the amateur except in special applications. No single maximum value should be used unless all other ratings can simultaneously be held within the maximum values. As an example, a tube may have a maximum plate-voltage rating of 2000, a maximum plate-current rating of 300 ma., and a maximum plate-power-input rating of 400 watts. Therefore, if the maximum plate voltage of 2000 is used, the plate current should be limited to 200 ma. (instead of 300 ma.) to stay within the maximum power-input rating of 400 watts.

SOURCES OF ELECTRODE VOLTAGES

Filament or Heater 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.

Plate Voltage

D.c. plate voltage for the operation of r.f. amplifiers is most often obtained from a transformer-rectifier-filter system (see power-supply chapter) designed to deliver the required plate voltage at the required current. However, batteries or other d.c.-generating devices are sometimes used in certain types of operation (see portable-mobile chapter).

Bias and Tube Protection

Several methods of obtaining bias are shown in Fig. 6-17. In A, bias is obtained by the voltage drop across a resistor in the grid d.c. return circuit when rectified grid current flows. The proper value of resistance may be determined by dividing the required biasing voltage by the d.c. grid current at which the tube will be operated. Then, so long as the r.f. driving voltage is adjusted so that the d.c. grid current is the recommended value, the biasing voltage will be the proper value. The tube is biased only when excitation is applied, since the voltage drop across the resistor depends upon grid-current flow. When excitation is removed, the bias falls to zero. At zero bias most tubes draw power far in excess of the plate-dissipation rating. So it is advisable to make provision for protecting the tube when excitation fails by accident, or by intent as it does when a preceding stage in a c.w. transmitter is keyed.

If the maximum c.w. ratings shown in the tube tables are to be used, the input should be cut to zero when the key is open. Aside from this, it is not necessary that plate current be cut off completely but only to the point where the rated

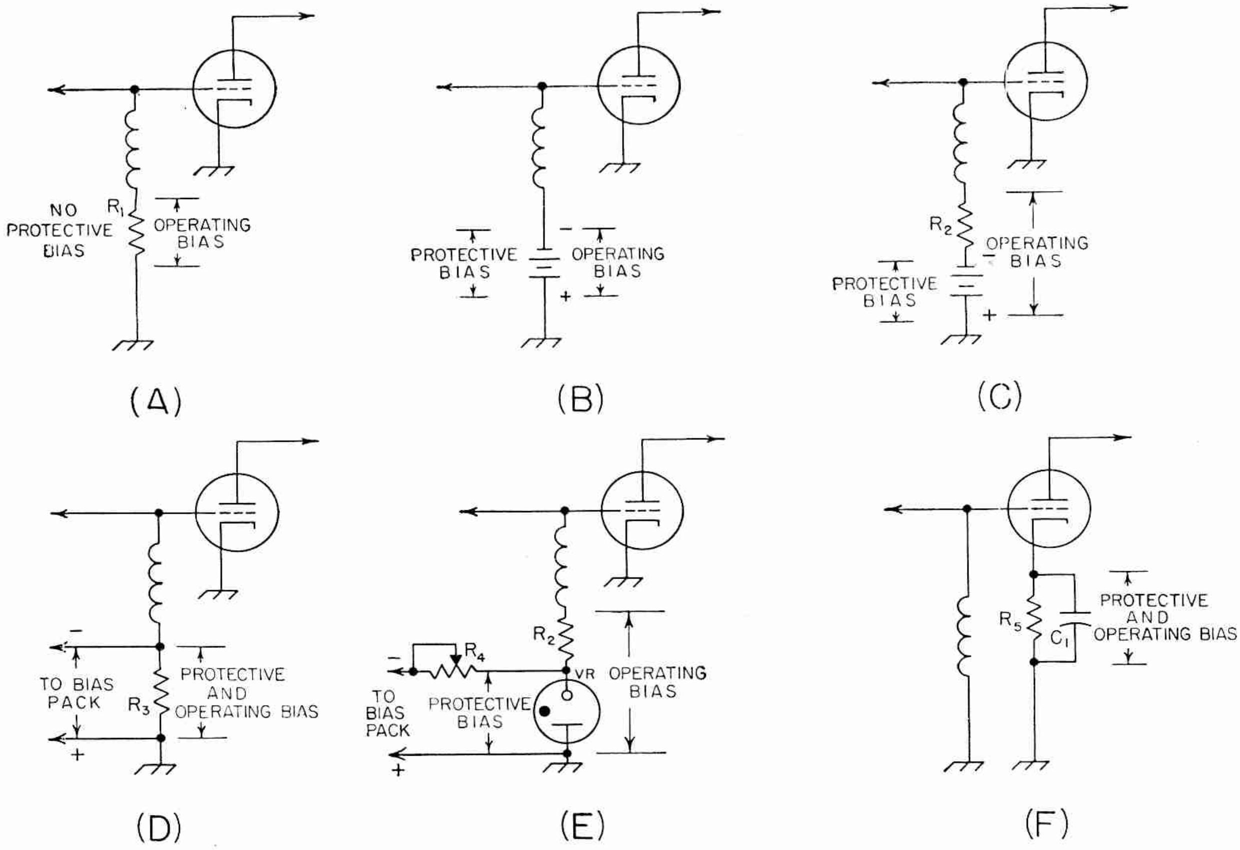


Fig. 6-17—Various systems for obtaining protective and operating bias for r.f. amplifiers. A—Grid-leak. B—Battery. C—Combination battery and grid leak. D—Grid leak and adjusted-voltage bias pack. E—Combination grid leak and voltage-regulated pack. F—Cathode bias.

dissipation is not exceeded. In this case plate-modulated phone ratings should be used for c.w. operation, however.

With triodes this protection can be supplied by obtaining all bias from a source of fixed voltage, as shown in Fig. 6-17B. It is preferable, however, to use only sufficient fixed bias to protect the tube and obtain the balance needed for operating bias from a grid leak, as in C. The grid-leak resistance is calculated as above, except that the fixed voltage is subtracted first.

Fixed bias may be obtained from dry batteries or from a power pack (see power-supply chapter). If dry batteries are used, they should be checked periodically, since even though they may show normal voltage, they eventually develop a high internal resistance. Grid-current flow through this battery resistance may increase the bias considerably above that anticipated. The life of batteries in bias service will be approximately the same as though they were subject to a drain equal to the grid current, despite the fact that the grid-current flow is in such a direction as to charge the battery, rather than to discharge it.

In Fig. 6-17F, bias is obtained from the voltage drop across a resistor in the cathode (or filament center-tap) lead. Protective bias is obtained by the voltage drop across R_5 as a result of plate (and screen) current flow. Since plate current must flow to obtain a voltage drop across the resistor, it is obvious that cut-off protective bias cannot be obtained. 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. For this reason, the use of cathode bias usually is limited to low-voltage tubes when the extra voltage is not difficult to obtain.

The resistance of the cathode biasing resistor R_5 should be adjusted to the value which will give the correct operating bias voltage 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. A disadvantage of this biasing system is that the cathode r.f. connection to ground depends upon a bypass capacitor. From the consideration of v.h.f. harmonics and stability with high-perveance tubes, it is preferable to make the cathode-to-ground impedance as close to zero as possible.

Screen Voltage

For c.w. operation, and under certain conditions of phone operation (see amplitude-modulation chapter), the screen may be operated from a power supply of the same type used for plate supply, except that voltage and current ratings should be appropriate for screen requirements. The screen may also be operated through a series resistor or voltage-divider from a source of higher voltage, such as the plate-voltage supply, thus making a separate supply for the screen unnecessary. Certain precautions are necessary, depending upon the method used.

It should be kept in mind that screen current varies widely with both excitation and loading. If the screen is operated from a fixed-voltage source, the tube should never be operated without plate voltage and load, otherwise the screen may be damaged within a short time. Supplying the screen through a series dropping resistor from a higher-voltage source, such as the plate supply, affords a measure of protection, since the resistor causes the screen voltage to drop as the current increases, thereby limiting the power drawn by the screen. However, with a resistor, the screen voltage may vary considerably with excitation, making it necessary to check the voltage at the screen terminal under actual operating conditions to make sure that the screen voltage is normal. Reducing excitation will cause the screen current to drop, increasing the voltage; increasing excitation will have the opposite effect. These changes are in addition to those caused by changes in bias and plate loading, so if a screen-grid tube is operated from a series resistor or a voltage divider, its voltage should be checked as one of the final adjustments after excitation and loading have been set.

An approximate value for the screen-voltage dropping resistor may be obtained by dividing the voltage *drop* required from the supply voltage (difference between the supply voltage and rated screen voltage) by the rated screen current in decimal parts of an ampere. Some further adjustment may be necessary, as mentioned above, so an adjustable resistor with a total resistance above that calculated should be provided.

Protecting Screen-Grid Tubes

Screen-grid tubes cannot be cut off with bias unless the screen is operated from a fixed-voltage supply. In this case the cut-off bias is approximately the screen voltage divided by the amplification factor of the screen. This figure is not always shown in tube-data sheets, but cut-off voltage may be determined from an inspection of tube curves, or by experiment.

When the screen is supplied from a series dropping resistor, the tube can be protected by the use of a clamper tube, as shown in Fig. 6-18. The grid-leak bias of the amplifier tube with excitation is supplied also to the grid of the clamper tube. This is usually sufficient to cut off the clamper tube. However, when excitation is removed, the clamper-tube bias falls to zero and it draws enough current through the screen dropping resistor usually to limit the input to the amplifier to a safe value. If complete screen-voltage cut-off is desired, a VR tube may be inserted in the screen lead as shown. The VR-tube voltage rating should be high enough so that it will extinguish when excitation is removed.

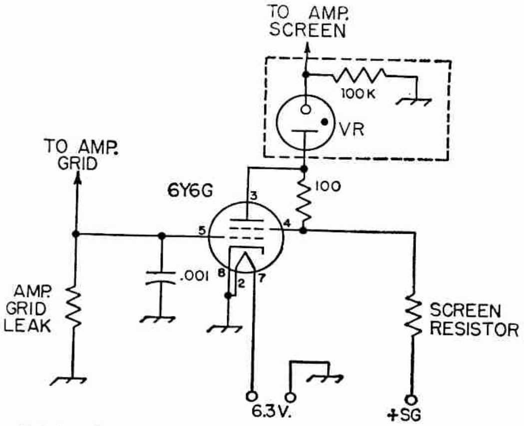


Fig. 6-18—Screen clamper circuit for protecting screen-grid power tubes. The VR tube is needed only for complete cut-off.

[illegible]
[illegible] Coupure complète

FEEDING EXCITATION TO THE GRID

The required r.f. driving voltage is supplied by an oscillator generating a voltage at the desired frequency, either directly or through intermediate amplifiers or frequency multipliers.

As explained in the chapter on vacuum-tube fundamentals, the grid of an amplifier operating under Class C conditions must have an exciting voltage whose peak value exceeds the negative biasing voltage over a portion of the excitation cycle. During this portion of the cycle, current will flow in the grid-cathode circuit as it does in a diode circuit when the plate of the diode is positive in respect to the cathode. This requires that the r.f. driver supply power. The power required to develop the required peak driving voltage across the grid-cathode impedance of the amplifier is the r.f. driving power.

The tube tables give approximate figures for the grid driving power required for each tube under various operating conditions. These figures, however, do not include circuit losses. In general, the driver stage for any Class C amplifier should be capable of supplying at least three times the driving power shown for typical operating conditions at frequencies up to 30 Mc., and from three to ten times at higher frequencies.

Since the d.c. grid current relative to the biasing voltage is related to the peak driving voltage, the d.c. grid current is commonly used as a convenient indicator of driving conditions. A driver adjustment that results in rated d.c. grid current when the d.c. bias is at its rated value, indicates proper excitation to the amplifier when it is fully loaded.

In coupling the grid input circuit of an amplifier to the output circuit of a driving stage the objective is to load the driver plate circuit so that the desired amplifier grid excitation is obtained without exceeding the plate-input ratings of the driver tube.

Driving Impedance

The grid-current flow that results when the grid is driven positive in respect to the cathode

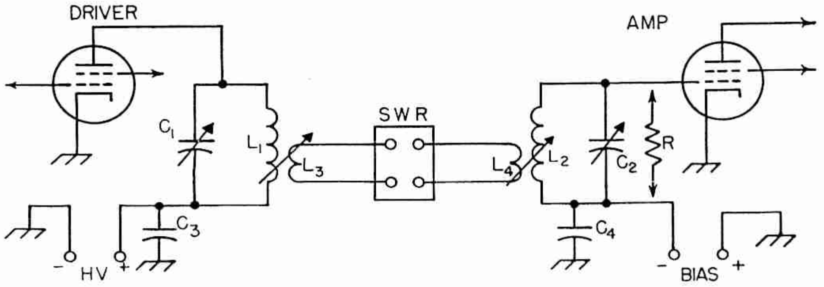


Fig. 6-19—Coupling excitation to the grid of an r.f. power amplifier by means of a low-impedance coaxial line.

C_1, C_3, L_1, L_3—See corresponding components in Fig. 6-10.

C_2—Amplifier grid tank capacitor—see text and Fig. 6-20 for capacitance, Fig. 6-34 for voltage rating.

C_4—0.-001-μf. disk ceramic.

L_2—To resonate at operating frequency with C_2. See *LC* chart inductance formula in electrical-laws chapter, or use ARRL *Lightning Calculator*.

L_4—Reactance equal to line impedance—see reactance chart and inductance formula in electrical-laws chapter, or use ARRL *Lightning Calculator*.

R is used to simulate grid impedance of the amplifier when a low-power s.w.r. indicator, such as a resistance bridge, is used. See formula in text for calculating value. Standing-wave indicator *SWR* is inserted only while line is made flat.

over a portion of the excitation cycle represents an average resistance across which the exciting voltage must be developed by the driver. In other words, this is the load resistance into which the driver plate circuit must be coupled. The approximate grid input resistance is given by:

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

For normal operation, the driving power and grid current may be taken from the tube tables.

Since the grid input resistance is a matter of a few thousand ohms, an impedance step-down is necessary if the grid is to be fed from a low-impedance transmission line. This can be done by the use of a tank as an impedance-transforming device in the grid circuit of the amplifier as shown in Fig. 6-19. This coupling 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 line is longer than a small fraction of a wave length, and if a s.w.r. bridge is available, the line is more easily handled by adjusting it as a matched transmission line.

Inductive Link Coupling with Flat Line

In adjusting this type of line, the object is to make the s.w.r. on the line as low as possible over as wide a band of frequencies as possible so that power can be transferred over this range without retuning. It is assumed that the output coupling considerations discussed earlier have been observed in connection with the driver plate circuit. So far as the amplifier grid circuit is concerned, the controlling factors are the *Q* of the tuned grid circuit, L_2C_2, (see Fig. 6-20) the inductance of the coupling coil, L_4, and the degree of coupling between L_2 and L_4. 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.

Assuming that the coupling is adjustable, start with a trial position of L_4 with respect to L_2, and adjust C_2 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_4 is too small. Maximum coupling, for a given degree of physi-

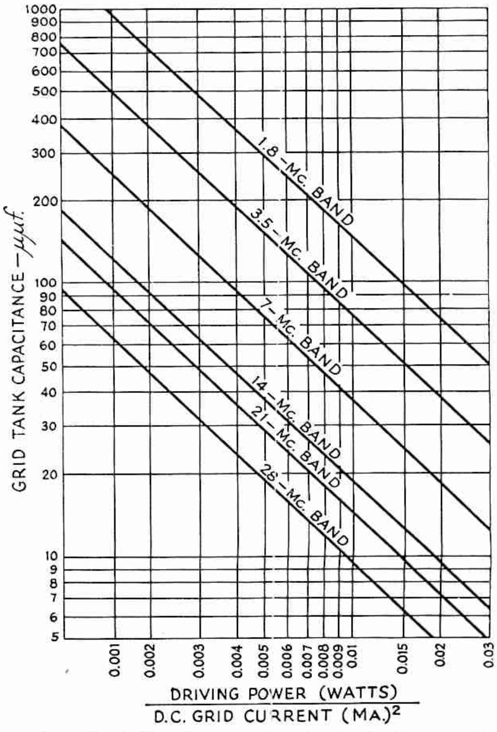


Fig. 6-20—Chart showing required grid tank capacitance for a Q of 12. To use, divide the driving power in watts by the square of the d.c. grid current in milliamperes and proceed as described under Fig. 6-9. Driving power and grid current may be taken from the tube tables. When a split-stator capacitor is used in a balanced grid circuit, the capacitance of *each section* may be half that shown.

cal coupling, will occur when the inductance of L_4 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_2 or the coupling between L_2 and L_4. 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_2 and correspondingly increasing L_2 to maintain resonance, and by tightening the coupling between L_2 and L_4, 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_2 and L_4 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_4 until a match is secured. The two coils should be tightly coupled.

When a resistance-bridge type s.w.r. indicator (see measuring-equipment section) 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. 6-19. In this case the amplifier tube *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. Coupling will be facilitated if the line is tuned as described under the earlier section on output coupling systems.

Link Feed with Unmatched Line

When the system is to be treated without regard to transmission-line effects, the link line must not offer appreciable reactance at the operating frequency. Any appreciable reactance will in effect reduce the coupling, making it impossible to transfer sufficient power from the driver to the amplifier grid circuit. Coaxial cables especially have considerable capacitance for even short lengths and it may be more desirable to use a spaced line, such as Twin-Lead, if the radiation can be tolerated.

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. Since the s.w.r. on the link line may be quite high, the line losses increase because of the greater current, the voltage increase may be sufficient to cause a breakdown in the insulation 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 tolerable at 3.5 Mc., but a length of a foot at 28 Mc. may be enough to cause serious effects on the functioning of the system.

Adjusting the coupling in such a system must necessarily be largely a matter of cut and try. If the line is short enough so as to have negligible reactance, the coupling between the two tank circuits will increase within limits by adding turns to the link coils, or by coupling the link coils more tightly, if possible, to the tank coils. If it is impossible to change either of these, a variable capacitor of 300 μμf. may be connected in series with or in parallel with the link coil at the driver end of the line, depending upon which connection is the most effective.

If coaxial line is used, the capacitor should be connected in series with the inner conductor. If the line is long enough to have appreciable reactance, the variable capacitor is used to resonate the entire link circuit.

As mentioned previously, the size of the link coils and the length of the line, as well as the size of the capacitor, will affect the resonant frequency and it may take an adjustment of all three before the capacitor will show a pronounced effect on the coupling.

When the system has been made resonant, coupling may be adjusted by varying the link capacitor.

Simple Capacitive Interstage Coupling

The capacitive system of Fig. 6-21A is the simplest of all coupling systems. (See Fig. 6-8 for filament-type tubes.) In this circuit, the plate tank circuit of the driver, C_1L_1, serves also as the grid tank of the amplifier. Although it is used more frequently than any other system, it is less flexible and has certain limitations that must be taken into consideration.

The two stages cannot be separated physically any appreciable distance without involving loss in transferred power, radiation from the coupling lead and the danger of feedback from this lead. Since both the output capacitance of the driver tube and the input capacitance of the amplifier are across the single circuit, it is sometimes difficult to obtain a tank circuit with a sufficiently low Q to provide an efficient circuit at the higher frequencies. The coupling can be varied by altering the capacitance of the coupling

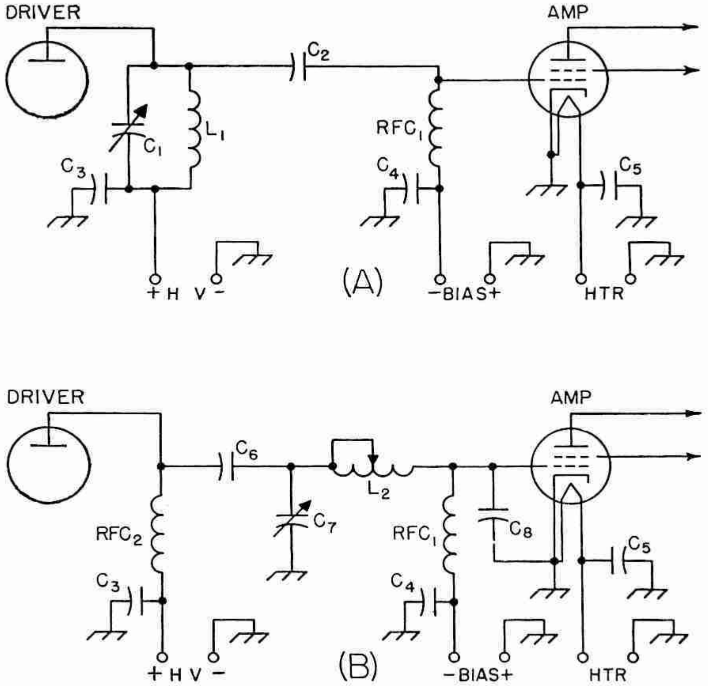


Fig. 6-21—Capacitive-coupled amplifiers. A—Simple capacitive coupling. B—Pi-section coupling.

C_1—Driver plate tank capacitor—see text and Fig. 6-9 for capacitance, Fig. 6-33 for voltage rating.
C_2—Coupling capacitor—50 to 150 $\mu\mu$f. mica, as necessary for desired coupling. Voltage rating sum of driver plate and amplifier biasing voltages, plus safety factor.
C_3—Driver plate bypass capacitor—0.001-μf. disk ceramic or mica. Voltage rating same as plate voltage.
C_4—Grid bypass—0.001-μf. disk ceramic.
C_5—Heater bypass—0.001-μf. disk ceramic.
C_6—Driver plate blocking capacitor—0.001-μf. disk ceramic or mica. Voltage rating same as C_2.
C_7—Pi-section input capacitor—see text referring to Fig. 6-12 for capacitance. Voltage rating—see Fig. 6-33A.
C_8—Pi-section output capacitor—100-$\mu\mu$f. mica. Voltage rating same as driver plate voltage plus safety factor.
L_1—To resonate at operating frequency with C_1. See *LC* chart and inductance formula in electrical-laws chapter, or use ARRL *Lightning Calculator.*
L_2—Pi-section inductor—See Fig. 6-12. Approx. same as L_1.
RFC_1—Grid r.f. choke—2.5-mh.
RFC_2—Driver plate r.f. choke—2.5 mh.

capacitor, C_2. The driver load impedance is the sum of the amplifier grid resistance and the reactance of the coupling capacitor in series, the coupling capacitor serving simply as a series reactor. The driver load resistance increases with a decrease in the capacitance of the coupling capacitor.

When the amplifier grid impedance is lower than the optimum load resistance for the driver, a transforming action is possible by tapping the grid down on the tank coil, but this is not recommended because it invariably causes an increase in v.h.f. harmonics and sometimes sets up a parasitic circuit.

So far as coupling is concerned, the *Q* of the circuit is of little significance. However, the other considerations discussed earlier in connection with tank-circuit *Q* should be observed.

Pi-Network Interstage Coupling

A pi-section tank circuit, as shown in Fig. 6-21B, may be used as a coupling device between screen-grid amplifier stages. The circuit is actually a capacitive coupling arrangement with the grid of the amplifier tapped down on the circuit by means of a capacitive divider. In contrast to the tapped-coil method mentioned previously, this system will be very effective in reducing v.h.f. harmonics, because the output capacitor, C_8, provides a direct capacitive shunt for harmonics across the amplifier grid circuit.

To be most effective in reducing v.h.f. harmonics, C_8 should be a mica capacitor connected directly across the tube-socket terminals. Tapping down on the circuit in this manner also helps to stabilize the amplifier at the operating frequency because of the grid-circuit loading provided by C_8. For the purposes both of stability and harmonic reduction, experience has shown that a value of 100 $\mu\mu$f for C_8 usually is sufficient. In general, C_7 and L_2 should have values approximating the capacitance and inductance used in a conventional tank circuit. A reduction in the inductance of L_2 results in an increase in coupling because C_7 must be increased to retune the circuit to resonance. This changes the ratio of C_7 to C_8 and has the effect of moving the grid tap up on the circuit. Since the coupling to the grid is comparatively loose under any condition, it may be found that it is impossible to utilize the full power capability of the driver stage. If sufficient excitation cannot be obtained, it may be necessary to raise the plate voltage of the driver, if this is permissible. Otherwise a larger driver tube may be required. As shown in Fig. 6-21B, parallel driver plate feed and amplifier grid feed are necessary.

● STABILIZING AMPLIFIERS

External Coupling

A straight amplifier operates with its input and output circuits tuned to the same frequency. Therefore, unless the coupling between these two circuits is brought to the necessary minimum, the amplifier will oscillate as a tuned-plate tuned-grid circuit. Care should be used in arranging components and wiring of the two circuits so that there will be negligible opportunity for coupling external to the tube itself. Complete shielding between input and output circuits usually is required. All r.f. leads should be kept as short as possible and particular attention should be paid to the r.f. return paths from plate and grid tank circuits to cathode. In general, the best arrangement is one in which the cathode (or filament center tap) connection to ground, and the plate tank circuit are on the same side of the chassis or other shielding. Then the "hot" lead from the grid tank (or driver plate tank) should be brought to the socket through a hole in the shielding. Then when the grid tank capacitor or bypass is grounded, a return path through the hole to cathode will be encouraged, since transmission-line characteristics are simulated.

A check on external coupling between input and output circuits can be made with a sensitive indicating device, such as the one diagrammed in Fig. 6-22. The amplifier tube is removed from its socket and if the plate terminal is at the socket, it should be disconnected. With the driver stage running and tuned to resonance, the indicator should be coupled to the output tank coil and the output tank capacitor tuned for any indication of r.f. feedthrough. Experiment with shielding and rearrangement of parts will show whether the isolation can be improved.

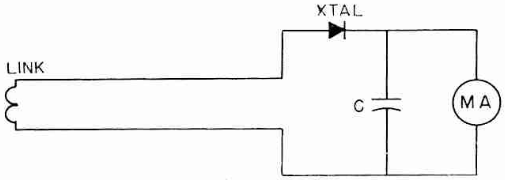


Fig. 6-22—Circuit of sensitive neutralizing indicator. *Xtal* is a 1N34 crystal detector, *MA* a 0–1 direct-current milliammeter and *C* a 0.001-μf. mica bypass capacitor.

Screen-Grid Neutralizing Circuits

The plate-grid capacitance of screen-grid tubes is reduced to a fraction of a micromicrofarad by the interposed grounded screen. Nevertheless, the power sensitivity of these tubes is so great that only a very small amount of feedback is necessary to start oscillation. To assure a stable amplifier, it is usually necessary to load the grid circuit, or to use a neutralizing circuit. A neutralizing circuit is one external to the tube that balances the voltage fed back through the grid-plate capacitance, by another voltage of opposite phase.

Fig. 6-23A shows how a screen-grid amplifier may be neutralized by the use of an inductive link line coupling the input and output tank circuits in proper phase. The two coils must be properly polarized. If the initial connection proves to be incorrect, connections to one of the link coils should be reversed. Neutralizing is adjusted by changing the distance between the link coils and the tank coils. In the case of capacitive coupling between stages, one of the link coils will be coupled to the plate tank coil of the driver stage.

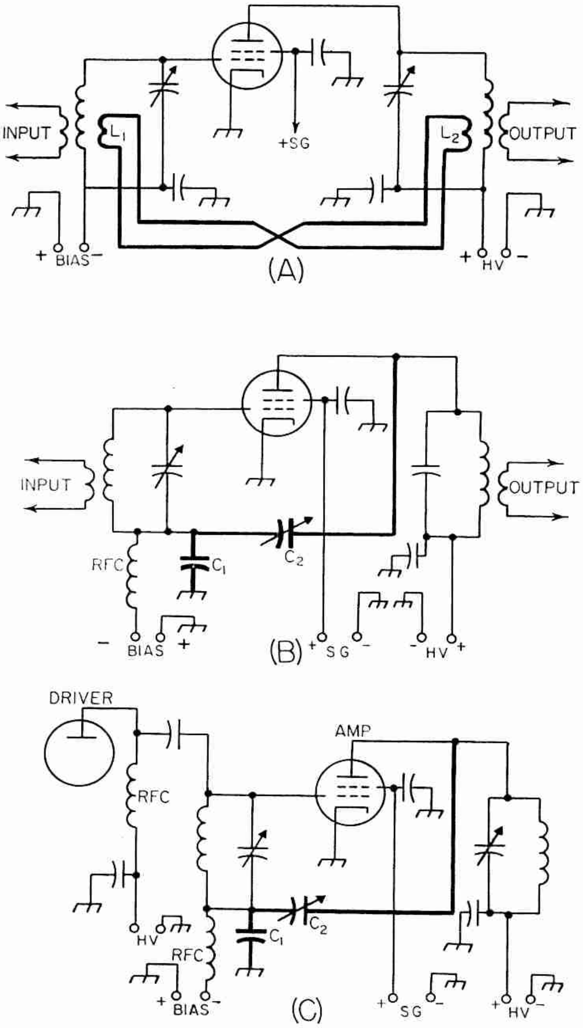


Fig. 6-23—Screen-grid neutralizing circuits. A—Inductive neutralizing. B—C—Capacitive neutralizing.

C_1—Grid bypass capacitor—approx. 0.001-μf. mica. Voltage rating same as biasing voltage in B, same as driver plate voltage in C.

C_2—Neutralizing capacitor—approx. 2 to 10 μμf.—see text. Voltage rating same as amplifier plate voltage for c.w., twice this value for plate modulation.

L_1, L_2—Neutralizing link—usually a turn or two will be sufficient.

A capacitive neutralizing system for screen-grid tubes is shown in Fig. 6-23B. C_2 is the neutralizing capacitor. The capacitance should be chosen so that at some adjustment of C_2,

$$\frac{C_2}{C_1} = \frac{\textit{Tube grid-plate capacitance} \text{ (or } C_{gp})}{\textit{Tube input capacitance} \text{ (or } C_{IN})}$$

The tube interelectrode capacitances C_{gp} and C_{IN} are given in the tube tables in the last chapter. The grid-cathode capacitance must include all strays directly across the tube capacitance, including the capacitance of the tuning-capacitor stator to ground. This may amount to 5 to 20 $\mu\mu$f. In the case of capacitance coupling, as shown in Fig. 6-23C, the output capacitance of the driver tube must be added to the grid-cathode capacitance of the amplifier in arriving at the value of C_2. If C_2 works out to an impractically large or small value, C_1 can be changed to compensate by using combinations of fixed mica capacitors in parallel.

Neutralizing Adjustment

The procedure in neutralizing is essentially the same for all types of tubes and circuits. The filament of the amplifier tube should be lighted and excitation from the preceding stage fed to the grid circuit. Both screen and plate voltages should be disconnected at the transmitter terminals.

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 carefully, bit by bit, the neutralizing capacitor or link coils until an r.f. indicator in the output circuit reads minimum.

The device shown in Fig. 6-22 makes a sensitive neutralizing indicator. 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. The plate tank capacitor should be readjusted for maximum reading after each change in neutralizing.

The grid-current meter may also be used as a neutralizing indicator. With plate and screen voltages removed as described above, there will be a change in grid current as the plate tank circuit is tuned through resonance. The neutralizing capacitor should be adjusted until this deflection is brought to a minimum. As a final adjustment, plate and screen voltages should be applied and the neutralizing capacitance adjusted to the point where minimum plate current, maximum grid current and maximum screen current occur simultaneously. An increase in grid current when the plate tank circuit is tuned slightly on the high-frequency side of resonance indicates that the neutralizing capacitance is too small. If the increase is on the low-frequency side, the neutralizing capacitance is too large. When neutralization is complete, there should be a slight decrease in grid current on either side of resonance.

Grid Loading

The use of a neutralizing circuit may often be avoided by loading the grid circuit if the driving stage has some power capability to spare. Loading by tapping the grid down on the grid tank coil (or the plate tank coil of the driver in the case of capacitive coupling), or by a resistor from grid to cathode is effective in stabilizing an amplifier, but either device may increase v.h.f. harmonics. The best loading system is the use of a pi-section filter, as shown in Fig. 6-21B. This circuit places a capacitance directly between grid and cathode. This not only provides the desirable loading, but also a very effective capacitive short for v.h.f. harmonics. A 100-$\mu\mu$f. mica capacitor for C_8, wired directly between tube terminals will usually provide sufficient loading to stabilize the amplifier.

V.H.F. Parasitic Oscillation

Parasitic oscillation in the v.h.f. range will take place in almost every r.f. power amplifier. To test for v.h.f. parasitic oscillation, the grid tank coil (or driver tank coil in the case of capacitive coupling) should be short-circuited with a clip lead. This is to prevent any possible t.g.t.p. oscillation at the operating frequency which might lead to confusion in identifying the parasitic. Any fixed bias should be replaced with a grid leak of 10,000 to 20,000 ohms. All load on the output of the amplifier should be disconnected. Plate and screen voltages 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 lamp in series with the primary of the plate transformer.

With power applied only to the amplifier under test, a search should be made by adjusting the input capacitor to several settings, including minimum and maximum, and turning the plate capacitor through its range for each of the grid-capacitor settings. Any grid current, or any dip or flicker in plate current at any point, indicates oscillation. This can be confirmed by an indicating absorption wavemeter tuned to the frequency of the parasitic and held close to the plate lead of the tube.

The heavy lines of Fig. 6-24A show the usual parasitic tank circuit, which resonates, in most cases, between 150 and 200 Mc. For each type of tetrode, there is a region, usually below the parasitic frequency, in which the tube will be self-neutralized. By adding the right amount of inductance to the parasitic circuit, its resonant frequency can be brought down to the frequency

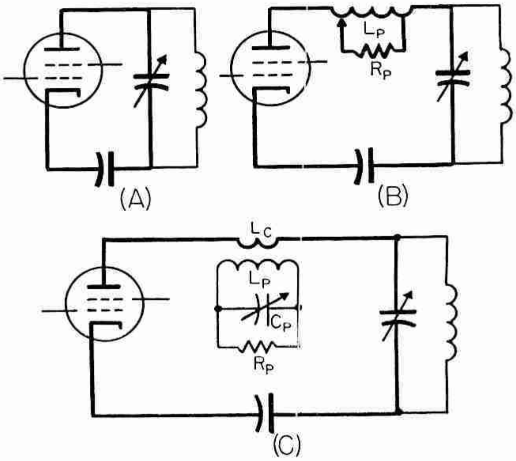


Fig. 6-24—A—Usual parasitic circuit. B—Resistive loading of parasitic circuit. C—Inductive coupling of loading resistance into parasitic circuit.

at which the tube is self-neutralized. However, the resonant frequency should not be brought down so low that it falls close to TV Channel 6 (88 Mc.). From the consideration of TVI, the circuit may be loaded down to a frequency not lower than 100 Mc. If the self-neutralizing frequency is below 100 Mc., the circuit should be loaded down to somewhere between 100 and 120 Mc. with inductance. Then the parasitic can be suppressed by loading with resistance, as shown in Fig. 6-24B. A coil of 4 or 5 turns, ¼ inch in diameter, is a good starting size. With the tank capacitor turned to maximum capacitance, the circuit should be checked with a g.d.o. to make sure the resonance is above 100 Mc. Then, with the shortest possible leads, a noninductive 100-ohm 1-watt resistor should be connected across the entire coil. The amplifier should be tuned up to its highest-frequency band and operated at low voltage. The tap should be moved a little at a time to find the minimum number of turns required to suppress the parasitic. Then voltage should be increased until the resistor begins to feel warm after several minutes of operation, and the power input noted. This input should be compared with the normal input and the *power* rating of the resistor increased by this proportion; i.e., if the power is half normal, the wattage rating should be doubled. This increase is best made by connecting 1-watt *carbon* resistors in parallel to give a resultant of about 100 ohms. As power input is increased, the parasitic may start up again, so power should be applied only momentarily until it is made certain that the parasitic is still suppressed. If the parasitic starts up again when voltage is raised, the tap must be moved to include more turns. So long as the parasitic is suppressed, the resistors will heat up only from the operating-frequency current.

Since the resistor can be placed across only that portion of the parasitic circuit represented by L_p, the latter should form as large a portion of the circuit as possible. Therefore, the tank and bypass capacitors should have the lowest possible inductance and the leads shown in heavy lines should be as short as possible and of the heaviest practical conductor. This will permit L_p to be of maximum size without tuning the circuit below the 100-Mc. limit.

Another arrangement that has been used successfully is shown in Fig. 6-24C. A small turn or two is inserted in place of L_p and this is coupled to a circuit tuned to the parasitic frequency and loaded with resistance. The heavy-line circuit should first be checked with a g.d.o. Then the loaded circuit should be tuned to the same frequency and coupled in to the point where the parasitic ceases. The two coils can be wound on the same form and the coupling varied by sliding one of them. Slight retuning of the loaded circuit may be required after coupling. Start out with low power as before, until the parasitic is suppressed. Since the loaded circuit in this case carries much less operating-frequency current, a single 100-ohm 1-watt resistor will often be sufficient and a 30-$\mu\mu$f. mica trimmer should serve as the tuning capacitor, C_p.

Low-Frequency Parasitic Oscillation

The screening of most transmitting screen-grid tubes is sufficient to prevent low-frequency parasitic oscillation caused by resonant circuits set up by r.f. chokes in grid and plate circuits. Should this type of oscillation (usually between 1200 and 200 kc.) occur, see paragraph under triode amplifiers.

● PARALLEL-TUBE AMPLIFIERS

The circuits for parallel-tube amplifiers are the same as for a single tube, similar terminals of the tubes being connected together. The grid impedance of two tubes in parallel is half that of a single tube. This means that twice the grid tank capacitance shown in Fig. 6-20 should be used for the same *Q*.

The plate load resistance is halved so that the plate tank capacitance for a single tube (Fig. 6-10) also should be doubled. The total grid current will be doubled, so to maintain the same grid bias, the grid-leak resistance should be half that used for a single tube. The required driving power is doubled. The capacitance of a neutralizing capacitor, if used, should be doubled and the value of the screen dropping resistor should be cut in half.

In treating parasitic oscillation, it may be necessary to use a choke in each plate lead, rather than one in the common lead. Input and output capacitances are doubled, which may be a factor in obtaining efficient operation at higher frequencies.

● PUSH-PULL AMPLIFIERS

Basic push-pull circuits are shown in Fig. 6-26C and D. Amplifiers using this circuit are cumbersome to bandswitch and consequently are not very popular below 30 Mc. However, since the push-pull configuration places tube input and output capacitances in series, the circuit is widely used at 50 Mc. and higher.

● TRIODE AMPLIFIERS

Circuits for triode amplifiers are shown in Fig. 6-26. Neglecting references to the screen, all of the foregoing information applies equally well to triodes. All triode straight amplifiers must be neutralized, as Fig. 6-26 indicates. From the tube tables, it will be seen that triodes require considerably more driving power than screen-grid tubes. However, they also have less power sensitivity, so that greater feedback can be tolerated without the danger of instability.

Low-Frequency Parasitic Oscillation

When r.f. chokes are used in both grid and plate circuits of a triode amplifier, the split-stator tank capacitors combine with the r.f. chokes to form a low-frequency parasitic circuit, unless the amplifier circuit is arranged to prevent it. In the circuit of Fig. 6-26B, the amplifier grid

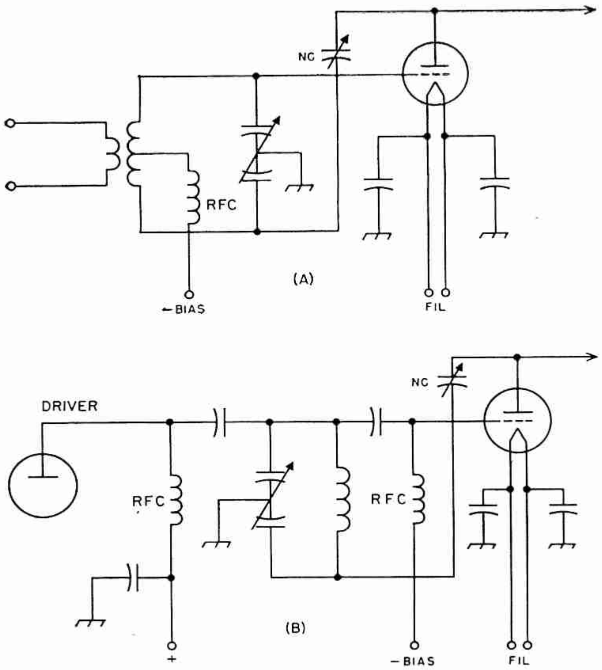


Fig. 6-25—When a pi-network output circuit is used with a triode, a balanced grid circuit must be provided for neutralizing. A—Inductive-link input. B—Capacitive input coupling.

is series fed and the driver plate is parallel fed. For low frequencies, the r.f. choke in the driver plate circuit is shorted to ground through the tank coil. In Figs. 6-26C and D, a resistor is substituted for the grid r.f. choke. This resistance should be at least 100 ohms. If any grid-leak resistance is used for biasing, it should be substituted for the 100-ohm resistor.

Triode Amplifiers with Pi-Network Output

Pi-network output tanks, designed as described earlier for screen-grid tubes, may also be used with triodes. However, in this case, a balanced input circuit must be provided for neutralizing. Fig. 6-25A shows the circuit when inductive-link input coupling is used, while B shows the circuit to be used when the amplifier is coupled capacitively to the driver. Pi-network circuits cannot be used in *both* input and output circuits, since no means is provided for neutralizing.

● GROUNDED-GRID AMPLIFIERS

Fig. 6-27A shows the input circuit of a grounded-grid triode amplifier. In configuration it is similar to the conventional grounded-cathode circuit except that the grid, instead of the cathode, is at ground potential. An amplifier of this type is characterized by a comparatively low input im-

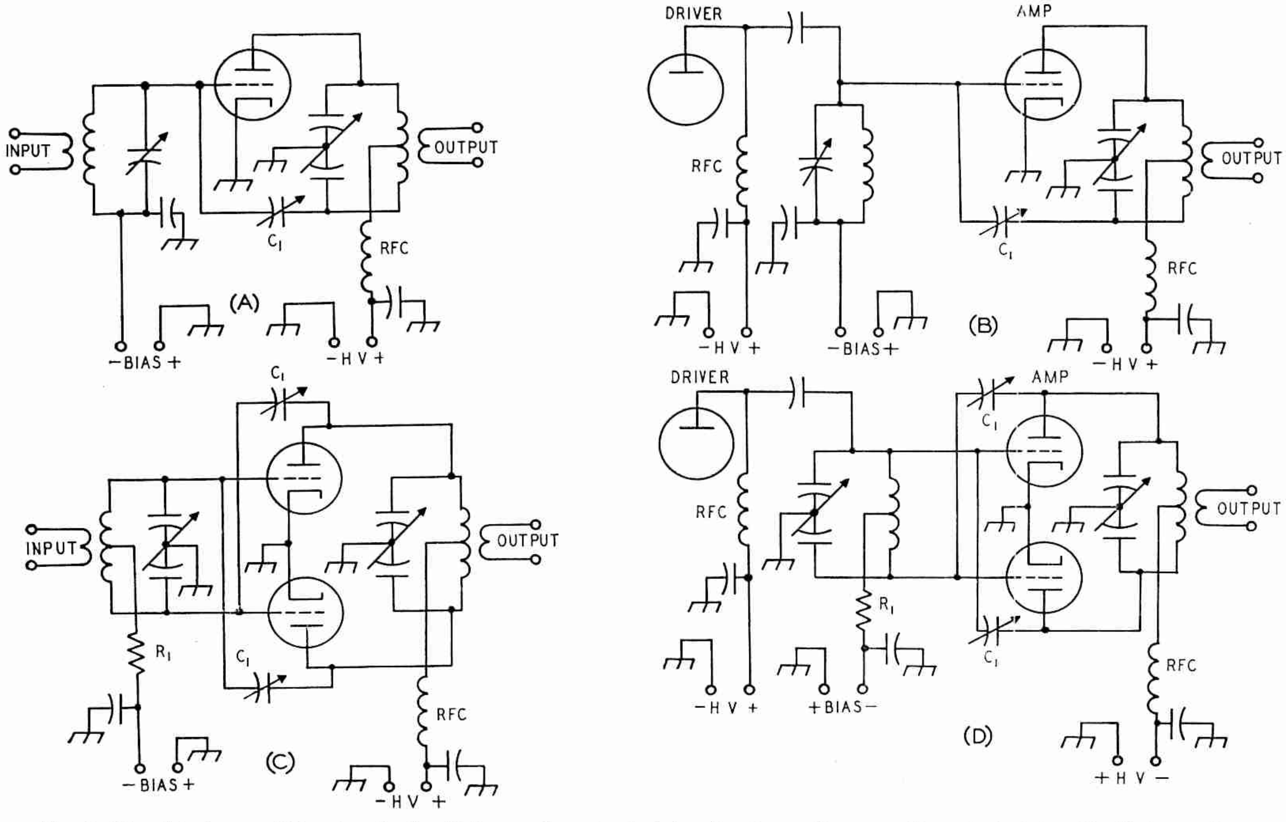


Fig. 6-26—Triode amplifier circuits. A—Link coupling, single tube. B—Capacitive coupling, single tube. C—Link coupling, push-pull. D—Capacitive coupling, push-pull. Aside from the neutralizing circuits, which are mandatory with triodes, the circuits are the same as for screen-grid tubes, and should have the same values throughout. The neutralizing capacitor, C_1, should have a capacitance somewhat greater than the grid-plate capacitance of the tube. Voltage rating should be twice the d.c. plate voltage for c.w., or four times for plate modulation, plus safety factor. The resistance R_1 should be at least 100 ohms and it may consist of part or preferably all of the grid leak. For other component values, see similar screen-grid diagrams.

Grounded-Grid Amplifiers

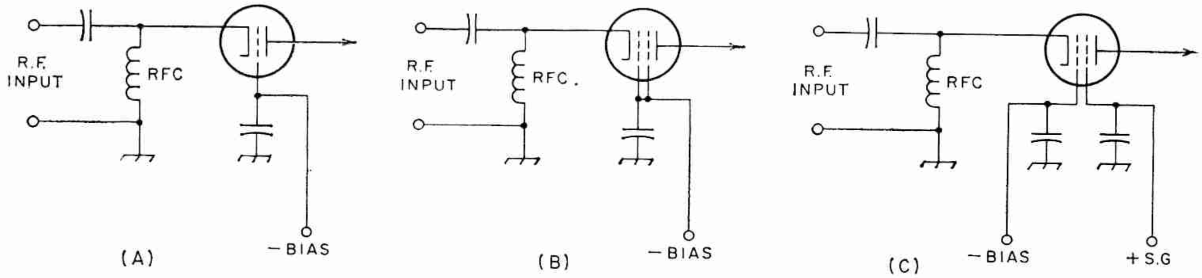


Fig. 6-27—A—Grounded-grid triode input circuit. B—Tetrode input circuit with grid and screen directly in parallel. C—Tetrode circuit with d.c. voltage applied to the screen. Plate circuits are conventional.

pedance and a relatively high driver-power requirement. The additional driver power is not consumed in the amplifier but is "fed through" to the plate circuit where it combines with the normal plate output power. The total r.f. power output is the sum of the driver and amplifier output powers less the power normally required to drive the tube in a grounded-cathode circuit.

Positive feedback is from plate to cathode through the plate-cathode, or plate-filament, capacitance of the tube. Since the grounded grid is interposed between the plate and cathode, this capacitance is very small, and neutralization usually is not necessary.

A disadvantage of the grounded-grid circuit is that the cathode must be isolated for r.f. from ground. This presents a practical difficulty, especially in the case of a filament-type tube whose filament current is large. Another disadvantage in plate-modulated phone operation is that the driver power fed through to the output is not modulated.

The chief application for grounded-grid amplifiers in amateur work at frequencies below 30 Mc. is in the case where the available driving power far exceeds the power that can be used in driving a conventional grounded-cathode amplifier.

D.c. electrode voltages and currents in grounded-grid triode-amplifier operation are the same as for grounded-cathode operation. Approximate values of driving power, driving impedance, and total power output in Class C operation can be calculated as follows, using information normally provided in tube data sheets. R.m.s. values are of the fundamental components:

$$E_p = \textit{r.m.s. value of r.f. plate voltage}$$

$$= \frac{\textit{d.c. plate volts} + \textit{d.c. bias volts} - \textit{peak r.f. grid volts}}{1.41}$$

$$I_p = \textit{r.m.s. value of r.f. plate current}$$

$$= \frac{\textit{rated power output watts}}{E_p}$$

$$E_g = \textit{r.m.s. value of grid driving voltage}$$

$$= \frac{\textit{peak r.f. grid volts}}{1.41}$$

$$I_g = \textit{r.m.s. value of r.f. grid current}$$

$$= \frac{\textit{rated driving power watts}}{E_g}$$

Then,

$$\textit{Driving power (watts)} = E_g\,(I_p + I_g)$$

$$\textit{Driving impedance (ohms)} = \frac{E_g}{I_g + I_p}$$

$$\textit{Power fed through from driver stage (watts)} = E_g I_p$$

$$\textit{Total power output (watts)} = I_p\,(E_g + E_p)$$

Screen-grid tubes are also used sometimes in grounded-grid amplifiers. In some cases, the screen is simply connected in parallel with the grid, as in Fig. 6-27B, and the tube operates as a high-μ triode. In other cases, the screen is by-passed to ground and operated at the usual d.c. potential, as shown at C. Since the screen is still in parallel with the grid for r.f., operation is very much like that of a triode except that the positive voltage on the screen reduces driver-power requirements. Since the information usually furnished in tube-data sheets does not apply to triode-type operation, operating conditions are usually determined experimentally. In general, the bias is adjusted to produce maximum output (within the tube's dissipation rating) with the driving power available.

Fig. 6-28 shows two methods of coupling a grounded-grid amplifier to the 50-ohm output of an existing transmitter. At A an L network is used, while a conventional link-coupled tank is shown at B. The values shown will be approximately correct for most triode amplifiers operating at 3.5 Mc. Values should be cut in half each time frequency is doubled, i.e., 250 $\mu\mu$f. and 7.5 μh. for 7 Mc., etc.

Filament Isolation

Since the filament or cathode of the grounded-grid amplifier tube operates at some r.f. potential above ground, it is necessary to isolate the filament from the power line. In the case of low-power tubes with indirectly heated cathodes, it is sometimes feasible to depend on the small capacitance existing between the heater and cathode, although it is preferable to provide additional isolation.

In Fig. 6-29, isolation is provided by a special low-capacitance filament transformer. RFC_1 carries only the cathode current. However, since transformers of this type are not generally avail-

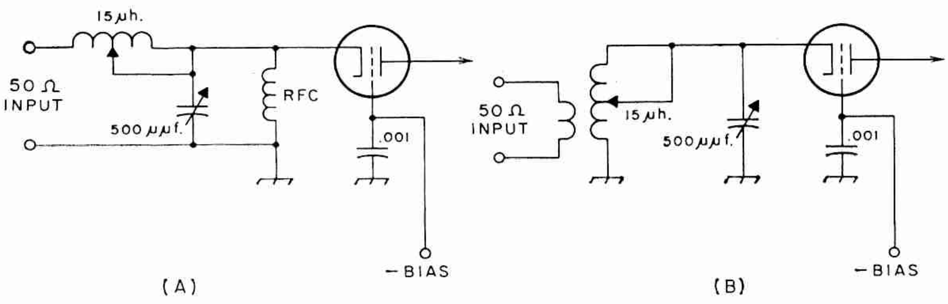


Fig. 6-28 — Two methods of coupling a low-impedance driver to a grounded-grid input. A—L network. B — Link-coupled tank circuit.

able, other means must usually be employed.

In Fig. 6-29B, chokes are used to isolate the filament from the filament transformer. The reactance of the chokes should be several times the input impedance of the amplifier and must be wound with conductor of sufficient size to carry the filament current. It is usually necessary to use a transformer delivering more than the rated filament voltage to compensate the voltage drop across the chokes. In Fig. 6-29C, r.f. chokes are placed in the primary side of the transformer. This reduces the current that the chokes must handle, but the filament transformer must be mounted so that it is spaced from the chassis and other grounded metal to minimize the capacitance of the transformer to ground. RFC_1 carries cathode current only.

In the case of the input circuit of Fig. 6-28B, it is sometimes feasible to wind the tank inductor with two conductors in parallel, and feed the filament voltage to the tube through the two conductors, as shown in Fig. 6-29D. This arrangement does not lend itself well to bandchanging, however.

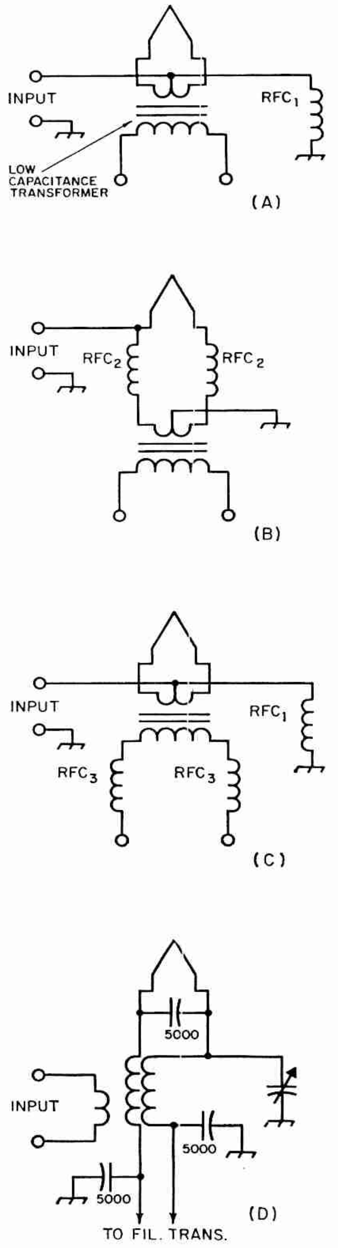
FREQUENCY MULTIPLIERS

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

Fig. 6-29—Methods of isolating filament from ground. A—Special low-capacitance filament transformer. B—R.f. chokes in filament circuit. C—R.f. chokes in transformer primary. D—Filament fed through input tank inductor.



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.

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.

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 neutralization. The link neutralizing system of Fig. 6-23A is convenient in such a contingency.

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. 6-30. 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, because there is a plate-current pulse for each cycle of the output frequency.

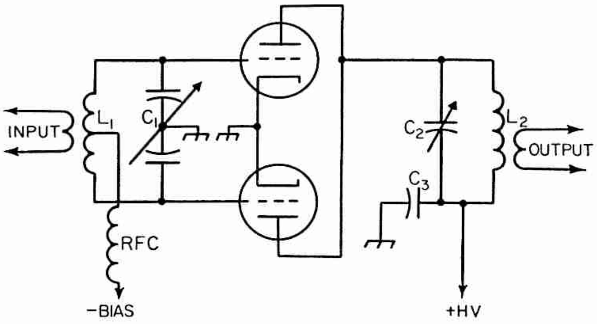


Fig. 6-30—Circuit of a push-push frequency multiplier for even harmonics.

C_1L_1 and C_2L_2—See text.

C_3—Plate bypass—0.001-μf. disk ceramic or mica. Voltage rating equal to plate voltage plus safety factor.

RFC—2.5-mh. r.f. choke.

This arrangement has an advantage in some applications. If the heater of one tube is turned off, 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.

The grid tank circuit is tuned to the frequency of the driving stage and should have the same constants as indicated in Fig. 6-20 for balanced grid circuits. The plate tank circuit is tuned to an even multiple of the exciting frequency, and should have the same values as a straight amplifier for the harmonic frequency (see Fig. 6-10), bearing in mind that the total plate current of both tubes determines the *C* to be used.

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, the plate tank circuit being tuned, of course, to the desired odd harmonic frequency.

● METERING

Fig. 6-31 shows how a voltmeter and milliammeter should be connected to read various voltages and currents. Voltmeters are seldom installed permanently, since their principal use is in preliminary checking. Also, milliammeters are not normally installed permanently in all of the positions shown. Those most often used are the ones reading grid current and plate current, or grid current and cathode current.

Milliammeters come in various current ranges. Current values to be expected can be taken from the tube tables and the meter ranges selected accordingly. To take care of normal overloads and pointer swing, a meter having a current range of about twice the normal current to be expected should be selected.

Meter Installation

Grid-current meters connected as shown in Fig. 6-31 and meters connected in the cathode circuit need no special precautions in mounting on the transmitter panel so far as safety is concerned. However, milliammeters having zero-adjusting screws on the face of the meter should be recessed behind the panel so that accidental contact with the adjusting screw is not possible, if the meter is connected in any of the other positions shown in Fig. 6-31. The meter can be mounted on a small subpanel attached to the front panel with long screws and spacers. The meter opening should be covered with glass or celluloid. Illuminated meters make reading easier. Reference should also be made to the TVI chapter of this *Handbook* in regard to wiring and shielding of meters to suppress TVI.

Meter Switching

Milliammeters are expensive items and there-

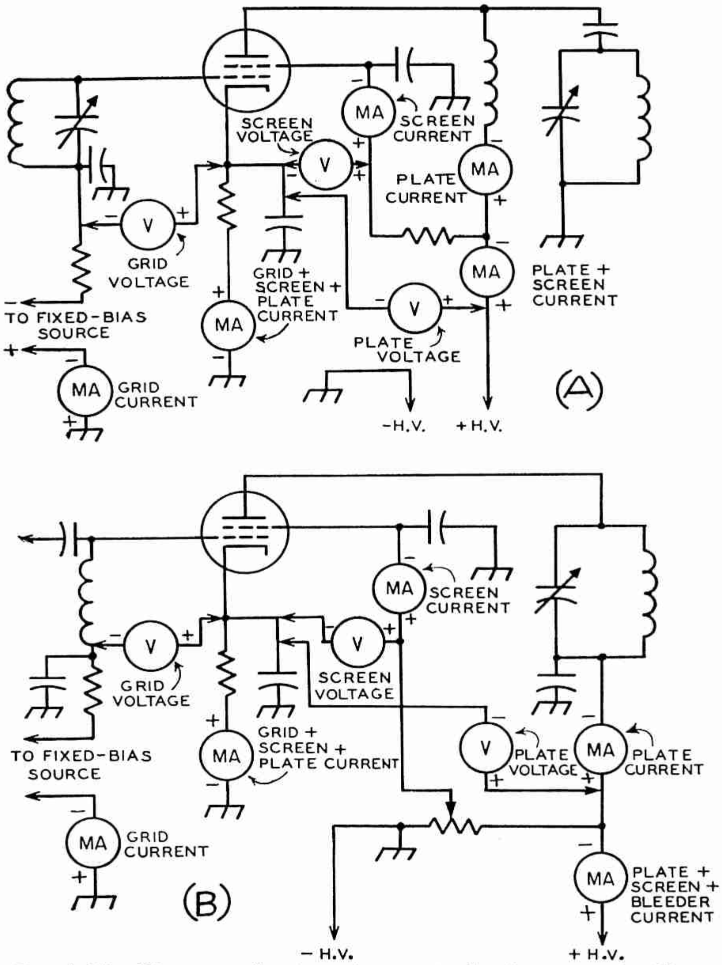


Fig. 6-31—Diagrams showing placement of voltmeter and milliammeter to obtain desired measurements. A—Series grid feed, parallel plate feed and series screen voltage-dropping resistor. B—Parallel grid feed, series plate feed and screen voltage divider.

fore it is seldom feasible to provide even grid-current and plate-current meters for all stages. The exciter stages in a multistage transmitter often do not require metering after initial adjustments. It is common practice to provide a meter-switching system by which a single milliammeter may be switched to read currents in as many circuits as desired. Such a meter-switching circuit is shown in Fig. 6-32. The resistors, *R*, are connected in the various circuits in place of the milliammeters shown in Fig. 6-31. 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.

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 Measurements.) Care should be taken to observe proper polarity in making the connections between the resistors and the switch.

AMPLIFIER ADJUSTMENT

Earlier sections in this chapter have dealt with the design and adjustment of input (grid) and output (plate) coupling systems, the stabilitization of amplifiers, and the methods of obtaining the required electrode voltages. Reference to these sections should be made as necessary in following a procedure of amplifier adjustment.

The objective in the adjustment of an intermediate amplifier stage is to secure adequate excitation to the following stage. In the case of the output or final amplifier, the objective is to obtain maximum power output to the antenna. In both cases, the adjustment must be consistent with the tube ratings as to voltage, current and dissipating ratings.

Adequate drive to a following amplifier is normally indicated when rated grid current in the following stage is obtained with the stage operating at rated bias, the stage loaded to rated plate current, and the driver stage tuned to resonance. In a final amplifier, maximum output is normally indicated when the output coupling is adjusted so that the amplifier tube draws rated plate current when it is tuned to resonance.

Resonance in the plate circuit is normally indicated by the dip in plate-current reading as the plate tank capacitor is tuned through its range. When the stage is unloaded, or lightly

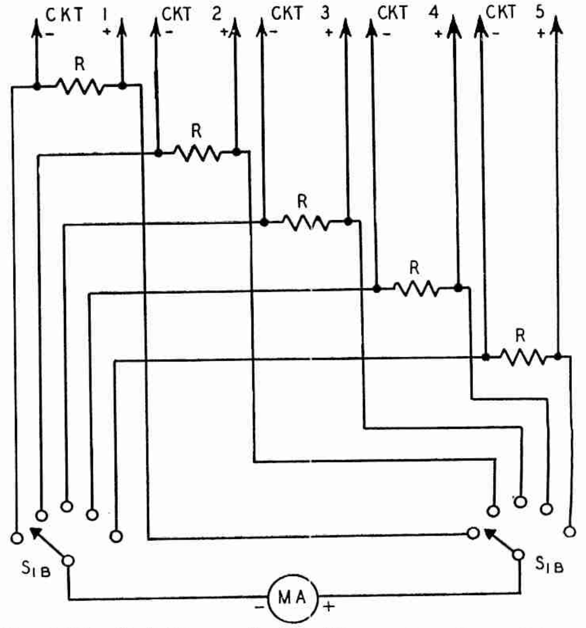


Fig. 6-32—Switching a single milliammeter. The resistors, *R*, should be 10 to 20 times the internal resistance of the meter; 47 ohms will usually be satisfactory. S_1 is a 2-section rotary switch. Its insulation should be ceramic for high voltages, and an insulating coupling should always be used between shaft and control.

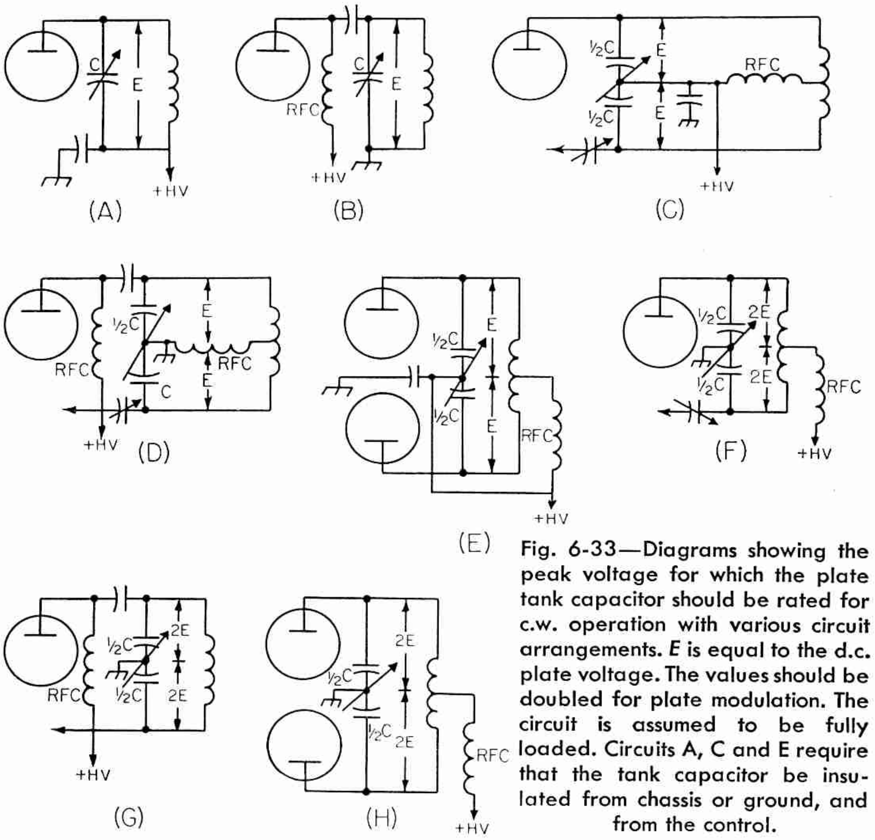


Fig. 6-33—Diagrams showing the peak voltage for which the plate tank capacitor should be rated for c.w. operation with various circuit arrangements. *E* is equal to the d.c. plate voltage. The values should be doubled for plate modulation. The circuit is assumed to be fully loaded. Circuits A, C and E require that the tank capacitor be insulated from chassis or ground, and from the control.

loaded, this dip in plate current will be quite pronounced. As the loading is increased, the dip will become less noticeable. See Fig. 6-4. However, in the case of a screen-grid tube whose screen is fed through a series resistor, maximum output may not be simultaneous with the dip in plate current. The reason for this is that the screen current varies widely as the plate circuit is tuned through resonance. This variation in screen current causes a corresponding variation in the voltage drop across the screen resistor. In this case, maximum output may occur at an adjustment that results in an optimum combination of screen voltage and nearness to resonance. This effect will seldom be observed when the screen is operated from a fixed-voltage source.

The first step in the adjustment of an amplifier is to stabilize it, both at the operating frequency by neutralizing it if necessary, and at parasitic frequencies by introducing suppression circuits.

If "flat" transmission-line coupling is used, the output end of the line should be matched, as described in this chapter for the case where the amplifier is to feed the grid of a following stage, or in the transmission-line section if the amplifier is to feed an antenna system. After proper match has been obtained, all adjustments in coupling should be made at the *input* end of the line.

Until preliminary adjustments of excitation have been made, the amplifier should be operated with filament voltage on and fixed bias, if it is required, but screen and plate voltages off. With the exciter coupled to the amplifier, the coupling to the driver should be adjusted until the amplifier draws rated grid current, or somewhat above the rated value. Then a load (the antenna grid of the following stage, or a dummy load) should be coupled to the amplifier.

With screen and plate voltages (preferably reduced) applied, the plate tank capacitor should be adjusted to resonance as indicated by a dip in plate current. Then, with full screen and plate voltages applied, the coupling to the load should be adjusted until the amplifier draws rated plate current. Changing the coupling to the load will usually detune the tank circuit, so that it will be necessary to readjust for resonance each time a change in coupling is made. An amplifier should not be operated with its plate circuit off resonance for any except the briefest necessary time, since the plate dissipation increases greatly when the plate circuit is not at resonance. Also, a screen-grid tube should not be operated without normal load for any appreciable length of time, since the screen dissipation increases.

It is normal for the grid current to decrease when plate voltage is applied, and to decrease again as the amplifier is loaded more heavily. As the grid current falls off, the coupling to the driver should be increased to maintain the grid current at its rated value.

● COMPONENT RATINGS AND INSTALLATION

Plate Tank-Capacitor Voltage

In selecting a tank capacitor 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 capacitor, 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 capacitor, to permit the use of a smaller capacitor with less plate spacing. Fig. 6-33 shows the peak voltage, in terms of d.c. plate voltage, to be expected across the tank capacitor 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.

The plate spacing to be used for a given peak voltage will depend upon the design of the variable capacitor, influencing factors being the mechanical construction of the unit, the insulation used and its placement in respect to intense fields, and the capacitor plate shape and degree of polish. Capacitor manufacturers usually rate their products in terms of the peak voltage between plates. Typical plate spacings are shown in the following table.

Typical Tank-Capacitor Plate Spacings

Spacing (In.)	*Peak Voltage*	*Spacing (In.)*	*Peak Voltage*	*Spacing (In.)*	*Peak Voltage*
0.015	1000	0.07	3000	0.175	7000
0.02	1200	0.08	3500	0.25	9000
0.03	1500	0.125	4500	0.35	11000
0.05	2000	0.15	6000	0.5	13000

Plate tank capacitors should be mounted as close to the tube as temperature considerations will permit to make possible the shortest capacitive path from plate to cathode. Especially at the higher frequencies where minimum circuit capacitance becomes important, the capacitor should be mounted with its stator plates well spaced from the chassis or other shielding. In circuits where the rotor must be insulated from ground, the capacitor should be mounted on ceramic insulators of size commensurate with the plate voltage involved and — most important of all, from the viewpoint of safety to the operator — a well-insulated coupling should be used between the capacitor shaft and the dial. *The section of the shaft attached to the dial should be well grounded.* This can be done conveniently through the use of panel shaft-bearing units.

Grid Tank Capacitors

In the circuit of Fig. 6-34, the grid tank capacitor should have a voltage rating approximately equal to the biasing voltage plus 20 per cent of the plate voltage. In the balanced circuit of B, the voltage rating of *each section* of the capacitor should be this same value.

The grid tank capacitor is preferably mounted with shielding between it and the tube socket for isolation purposes. It should, however, be mounted close to the socket so that a short lead can be passed through a hole to the socket. The rotor ground lead or bypass lead should be run directly to the nearest point on the chassis or other shielding. In the circuit of Fig. 6-34A, the same insulating precautions mentioned in connection with the plate tank capacitor should be used.

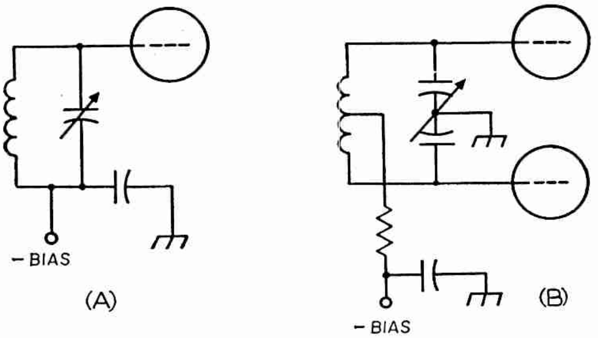


Fig. 6-34—The voltage rating of the grid tank capacitor in A should be equal to the biasing voltage plus about 20 per cent of the plate voltage.

Plate Tank Coils

The inductance of a manufactured coil usually is based upon the highest plate-voltage/plate-current ratio likely to be used at the maximum power level for which the coil is designed. Therefore in the majority of cases, the capacitance shown by Figs. 6-9 and 6-20 will be greater than that for which the coil is designed and turns must be removed if a Q of 10 or more is needed. 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 (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.

Tank coils should be mounted at least their diameter away from shielding to prevent a marked loss in Q. Except perhaps at 28 Mc., it is not important that the coil be mounted quite close to the tank capacitor. Leads up to 6 or 8 inches are permissible. It is more important to keep the tank capacitor as well as other components out of the immediate field of the coil. For this reason, it is preferable to mount the coil so that its axis is parallel to the capacitor shaft, either alongside the capacitor or above it.

There are many factors that must be taken into consideration in determining the size of wire that should be used in winding a tank coil. The considerations of form factor and wire size that will produce a coil of minimum loss are often of less importance in practice than the coil size that will fit into available space or that will handle the required power without excessive heating. This is particularly true in the case of screen-grid tubes where the relatively small driving power required can be easily obtained even if the losses in the driver are quite high. It may be considered preferable to take the power loss if the physical size of the exciter can be kept down by making the coils small.

The accompanying table shows typical conductor sizes that are usually found to be adequate for various power levels. For powers under 25 watts, the minimum wire sizes shown are largely a matter of obtaining a coil of reasonable Q. So far as the power is concerned, smaller wire could be used.

Wire Sizes for Transmitting Coils

Power Input (Watts)	*Band (Mc.)*	*Wire Size*
1000	28–21	6
	14–7	8
	3.5–1.8	10
500	28–21	8
	14–7	12
	3.5–1.8	14
150	28–21	12
	14–7	14
	3.5–1.8	18
75	28–21	14
	14–7	18
	3.5–1.8	22
25 or less*	28–21	18
	14–7	24
	3.5–1.8	28

* Wire size limited principally by consideration of Q.

Space-winding the turns invariably will result in a coil of higher Q, especially at frequencies above 7 Mc., and a form factor in which the turns spacing results in a coil length between 1 and 2 times the diameter is usually considered satisfactory. Space winding is especially desirable at the higher power levels because the heat developed is dissipated more readily. The power lost in a tank coil that develops appreciable heat at the higher-power levels does not usually represent a serious loss percentagewise. A more serious consequence, especially at the higher frequencies, is that coils of the popular “air-wound’ type supported on plastic strips may deform. In this case, it may be necessary to use wire (or copper tubing) of sufficient size to make the coil self-supporting. Coils wound on tubular forms of ceramic or mica-filled bakelite will also stand higher temperatures.

Plate-Blocking and Bypass Capacitors

Plate-blocking capacitors should have low inductance; therefore capacitors of the mica or ceramic type are preferred. For frequencies between 3.5 and 30 Mc., a capacitance of 0.001 is commonly used. The voltage rating should be 25 to 50% above the plate-supply voltage (twice this rating for plate modulation).

Small disk ceramic capacitors (approximately ¼ inch in diameter) are to be preferred as bypass capacitors, since when they are applied correctly (see TVI chapter), they are series resonant in the TV range and therefore are an important measure in filtering power-supply leads. Capacitors of this type are rated at 600 to 1000 volts. At higher voltages, disk ceramics with higher-voltage ratings, or capacitors of the TV “doorknob” type are recommended. Voltage ratings of bypass capacitors should be similar to those for blocking capacitors.

R. F. Chokes

The characteristics of any r.f. choke will vary with frequency, from characteristics resembling those of a parallel-resonant circuit, of high impedance, to those of a series-resonant circuit, where the impedance is lowest. In between these extremes, the choke will show varying amounts of inductive or capacitive reactance.

In series-feed circuits, these characteristics are of relatively small importance because, in a correctly operating circuit, the r.f. voltage across the choke is negligible. In a parallel-feed circuit, however, the choke is shunted across the tank circuit, and is subject to the full tank r.f. voltage. If the choke does not present a sufficiently high impedance, enough power will be absorbed by the choke to cause it to burn out. With chokes of the usual type, wound with small wire for compactness, a relatively small amount of power loss in the choke will cause excessive heating.

To avoid this, the choke must have a sufficiently high reactance to be effective at the lowest frequency, and yet have no series resonances near the higher-frequency bands. The design of a choke that meets requirements over a range as wide as 3.5 to 30 Mc. at the higher voltages is quite critical.

Universal pie-wound chokes of the “receiver” type (2.5 mh., 125 ma.) are usually satisfactory if the plate voltage does not exceed 750. For higher voltages, a single-layer solenoid-type choke of correct design has been found satisfactory. The National type R-175A and Raypar RL-100, RL-101 and RL-102 are representative manufactured types. An example of a satisfactory home-made choke for voltages up to at least 3000 consists of 112 turns of No. 26 wire, spaced to a length of 3⅞ inches on a 1-inch ceramic form (Centralab stand-off insulator, type X3022H). A ceramic form is advisable from the consideration of temperature. This choke has only one series resonance (near 24 Mc.), and exhibits an equivalent parallel resistance of 0.25 megohm or more in all of the amateur bands from 80 through 10.

Since the characteristics of a choke will be affected by any metal in its field, it should be checked when mounted in the position in which it is to be used, or in a temporary set-up simulating the same conditions. The plate end of the choke should not be connected, but the power-supply end should be connected directly, or by-passed, to the chassis. The g.d.o. should be coupled as close to the ground end of the choke as possible. Series resonances, indicating the frequencies of greatest loss, should be checked with the choke short-circuited with a short piece of wire. Parallel resonances, indicating frequencies of least loss are checked with the short removed.

A Three-Band Oscillator Transmitter for the Novice

The novice transmitter shown in Figs. 6-35–6-38, inclusive, is easy to build and get working. It is a crystal-controlled, one-tube oscillator capable of running at 30 watts input on the 3.5-, 7, and 21 Mc. Novice bands. A special feature of the transmitter is a built-in keying monitor which permits the operator to listen to his own sending.

Regulated voltage is used on the screen of the oscillator. This minimizes frequency shift of the oscillator with keying, which is the cause of chirp. In addition, a small amount of cathode bias (R_4) is used on the oscillator. This also tends to improve the keying characteristics in a cathode-keyed simple-oscillator transmitter.

Circuit Details

The oscillator circuit used is the grid-plate type, and the tube is a 6DQ6A pentode. The power output is taken from the plate circuit of the tube. On 80 meters, an 80-meter crystal is needed. On 40, either 80- or 40-meter crystals can be used, although slightly more output will be obtained by using 40-meter crystals. To operate on 15 meters, a 40-meter crystal is used.

The tank circuit is a pi network. The plate tank capacitor is the variable C_6, and the tank inductance is L_2L_3. C_8 is a two-section variable, approximately 365 $\mu\mu$f. per section, with the stators connected together to give a total capacitance of about 730 $\mu\mu$f. This range of capacitance is adequate for coupling to 50 or 75 ohms on 7 and 21 Mc. When operating on 3.5 Mc., an additional 1000 $\mu\mu$f. (C_7) is added to furnish the needed range of capacitance. L_1 and R_2 are essential for suppressing v.h.f. parasitic oscillations.

The keying-monitor circuit uses a neon bulb (type NE-2) audio-frequency oscillator connected to the cathode of the 6DQ6A at the key jack, J_1. The headphones are plugged into J_2, a jack mounted on the back of the transmitter chassis. Another jack, J_3, is used as a terminal for the leads that go to the headphone jack on the receiver.

Power Supply

The power supply uses a 5U4G in a full-wave circuit. A capacitor-input filter is used and the output voltage is approximately 370 volts with a cathode current of 90 milliamperes. A 0–150 milliammeter reads cathode current. The screen and grid currents are approximately 4 ma. when the oscillator is loaded.

Construction

All of the components, including the power supply, are mounted on a 2 × 7 × 13-inch aluminum chassis that is in turn enclosed in a 7 × 9 × 15-inch aluminum box. (Premier AC-1597). One of the removable covers of the box is used as the front panel, as shown in Fig. 6-35. The box has a ½-inch lip around both openings, so the bottom edge of the chassis should be placed one inch from the bottom of the panel. The sides of the chassis are also one inch from the sides of the panel. The chassis is held to the panel by S_2, J_1, and the mounting screws for the crystal socket, so both the front edge of the chassis and the panel must be drilled alike for these components. S_1, at the left in the front view, is one inch from the edge of the chassis (that is, two inches from the edge of the panel) and centered vertically on the chassis edge. Thus it is one inch from the bottom of the chassis edge and two inches from the bottom edge of the panel. The hole for J_1 is centered on the chassis edge and the holes for the crystal socket are drilled at the right-hand end of the chassis to correspond with the position of S_1 at the left.

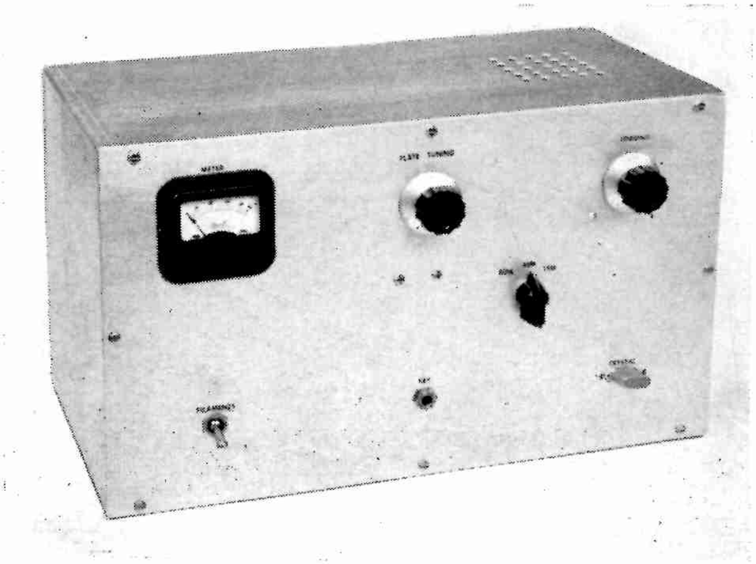


Fig. 6-35—This 30-watt three-band Novice transmitter is enclosed in a 7 × 9 × 15-inch aluminum box. A group of ¼-inch-diameter holes should be drilled in the top of the box over the oscillator tube, as shown, to provide ventilation. A similar set of holes should be drilled in the back cover behind the oscillator circuit.

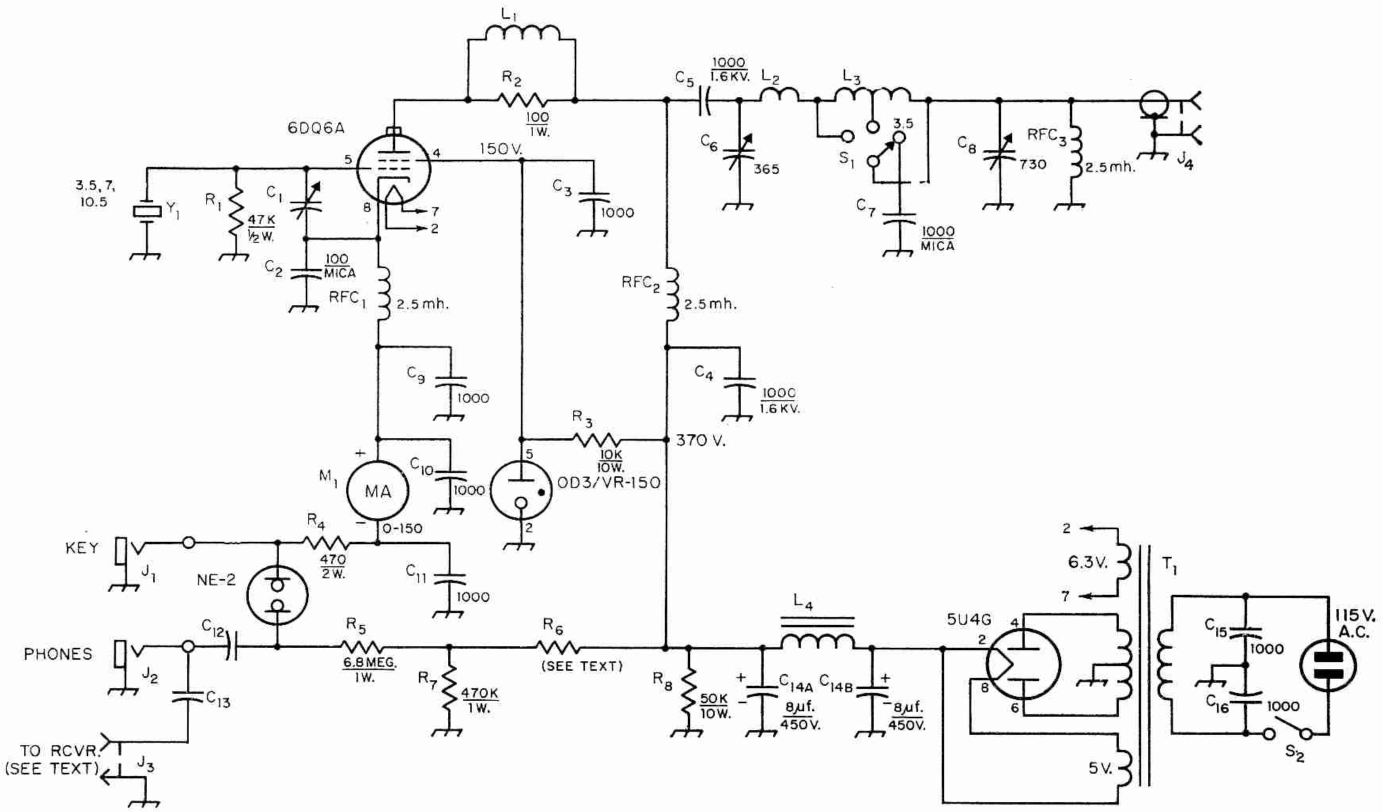


Fig. 6-36—Circuit diagram of the three-band transmitter. Unless otherwise specified, capacitances are in μμf. Resistances are in ohms (K= 1000).

C_1—3-30-μμf. trimmer.
C_2—100-μμf. mica.
C_3, C_9, C_{10}, C_{11}, C_{15}, C_{16}—0.001-μf. disk ceramic.
C_4, C_5—0.001-μf. 1600-volt disk ceramic.
C_6—365-μμf. variable capacitor, single section, broadcast-replacement type.
C_7—0.001-μf. 600-volt mica.
C_8—365-μμf. variable capacitor, dual section, broadcast-replacement type.
C_{12}—500-μμf. mica or ceramic.
C_{13}—0.01-μf. disk ceramic.
C_{14}—8/8-μf. 450-volt dual electrolytic capacitor.
J_1, J_2—Open-circuit phone jack.
J_3—Phono jack, RCA type.
J_4—Coaxial chassis connector, SO-239.
L_1—10 turns No. 18 wire space-wound on R_2.
L_2—6 turns No. 16 wire, 8 turns per inch, 1¼ inches diam. (B & W 3018).
L_3—23 turns No. 16 wire, 8 turns per inch, 1¼ inches diam. (B & W 3018). The 7-Mc. tap is 18 turns from the junction of L_2 and L_3.
L_4—8-h. 150-ma. filter choke (Thordarson 20C54).
M_1—0-150 ma. (Shurite 950).
R_1–R_8 inc.—As specified.
RFC_1, RFC_2, RFC_3—2.5-mh. r.f. choke (National R-50 or or similar).
S_1—Single-pole 3-position switch (Centralab 1461).
S_2—Single-pole single-throw toggle switch.
T_1—Power transformer: 360–0–360 volts, 120 ma.; 6.3 volts, 3.5 amp.; 5 volts, 3 amp (Stancor PM-8410).
Y_1—Crystal (see text).

There is nothing critical about the placement of the meter or the shafts for C_6, C_8 and S_1. As shown in Fig. 6-38, C_6 is mounted directly above J_1 and approximately two inches from the top of the panel. C_8 similarly is above the crystal socket and on the same horizontal line as C_6. S_1 is about at the middle of the square formed by these four components.

The holes on the rear edge of the chassis for the coaxial connector J_4, phone jack J_2, receiver connector J_3, and for the a.c. cord are drilled at the same height as those on the front edge. Access holes should be cut in the rear cover of the box at the corresponding positions; these holes may be large enough to clear the components, but not larger than is necessary for this purpose. The cover fits tightly against the rear edge of the chassis and thus maintains the shielding for preventing radiation of harmonics in the television bands. However, it is advisable to fasten the cover to the chassis edge with a few sheet-metal screws, in order to insure good electrical contact.

There are several different types of broadcast-replacement variable capacitors on the market. Some of these have holes tapped in the front of the frame, and this type can be mounted directly on the panel using machine screws and spacers. Others have mounting holes only in the bottom. In this case, the capacitor can be mounted on a pair of L-shaped brackets made from strips of aluminum.

Both L_2 and L_3 are supported by their leads. One end of L_3 is connected to the stator of C_8 and the other end is connected to a junction on top of a one-inch-long steatite stand-off insulator. L_2 has one end connected to the stator of C_6 and the other end to one of the terminals on S_1.

Fig. 6-37—Rear view of the transmitter showing the placement of components above chassis. The loading capacitor, C_8, is at the left, L_3 is the vertical coil and L_2 the horizontal one. Rubber grommets are used to prevent chafing and to furnish additional insulation on the leads coming from below chassis.

The voltage-dividing network consisting of R_6 and R_7 provides the correct voltage for operating the keying monitor, R_6 is 1.65 megohms, a value obtained by using two 3.3-megohm 1-watt resistors in parallel. These resistors and other small components may be mounted on standard bakelite tie points.

Adjustment and Testing

When the unit is ready for testing, a 15- or 25-watt electric light will serve as a dummy load. One side of the lamp should be connected to the output lead and the other side to chassis ground. A crystal appropriate for the band to be used should be plugged into the crystal socket, and a key connected to the key jack. S_1 should be set to the proper band. S_2 may then be closed and the transmitter allowed to warm up.

Set C_8 at maximum capacitance (plates completely meshed) and close the key. Quickly tune C_6 to resonance, as indicated by a dip in the cathode-current reading. Gradually decrease the capacitance of C_8, while retouching the tuning of C_6 as the loading increases. Increased loading will be indicated by increasing lamp brightness and by larger values of cathode current. Tune for maximum lamp brilliance. The cathode current should read between 90 and 100 milliamperes when the oscillator is fully loaded.

C_1 should be adjusted for the best keying characteristics consistent with reasonably good power output. It is not advisable to attempt to adjust C_1 with a lamp dummy load, since the lamp resistance will change during the heating and cooling that take place during keying, and this will affect the keying characteristic of the oscillator. Use a regular antenna, with or without an antenna coupler or matching network as the antenna system may require, and listen to the keying on the station receiver. Remove the antenna from the receiver to prevent overloading, and adjust the r.f. gain control for a signal level comparable with that at which signals on that band are normally heard. Further details on checking keying will be found in the chapter on keying and break-in.

(Originally described in *QST* December, 1957.)

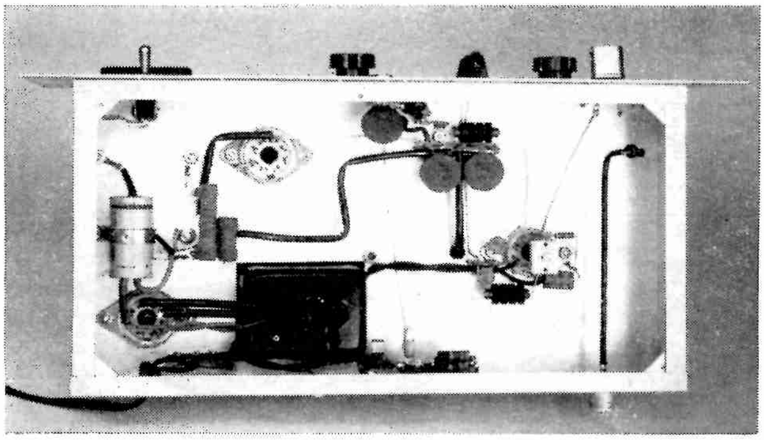


Fig. 6-38—Below-chassis view. Power-supply components are mounted in the left-hand side and the oscillator section is at the right-hand side. Mounted on the back wall of the chassis is the keying monitor. Although not visible in this view, the monitor components are mounted on a four-terminal tie point.

A One-Tube 50-Watt Transmitter

The transmitter shown in Figs. 6-39 and 6-41 is similar in some respects to the one described previously. However, it demonstrates a different type of construction and will handle more power. For simplicity, operation is confined to two bands — 80 and 40 meters.

The circuit is shown in Fig. 6-40. The single 6146 is used in a Colpitts-type crystal-oscillator circuit. The dial lamp I_1 serves as an indicator of r.f. crystal current and will also act as a fuse in case the crystal current becomes sufficient to endanger the crystal. (A crystal will fracture if the current through the crystal is sufficient to cause excessive heating.)

The output circuit, consisting of C_2, L_1 and C_4, is a pi network designed to feed a low-impedance (50–75-ohm) load. The band switch S_1 shorts out a portion of the coil for 40-meter operation and adds C_3 in parallel with C_4 for 80-meter output.

One of the functions of the r.f. choke RFC_4 is that of a safety device. Should the 1000-$\mu\mu$f. 1200-volt blocking capacitor break down, high voltage would be fed to the antenna or transmission line — a dangerous situation for the operator. The choke provides a d.c. short to ground should this occur, although it has no effect on the normal operation of the transmitter. The choke also makes it possible to use capacitors with a lower break-down voltage rating at C_2 and C_4.

The meter M_1 and the key are in the cathode circuit. Screen voltage is obtained from a voltage divider consisting of R_1 and R_2. R_1 consists of three 33,000-ohm 1-watt resistors connected in parallel, and R_2 is two 100,000-ohm resistors in parallel. If desired, 10,000-ohm and 50,000-ohm 10-watt resistors can be used instead.

Power Supply

A power supply delivering approximately 400 volts is included. The supply uses a 5U4GA or 5R4GY rectifier and a capacitive-input filter. The 100,000-ohm bleeder resistance across the output of the supply (shown in Fig. 6-40 as 100K, 5 watts) is made up of three 33,000-ohm, 2-watt resistors in series.

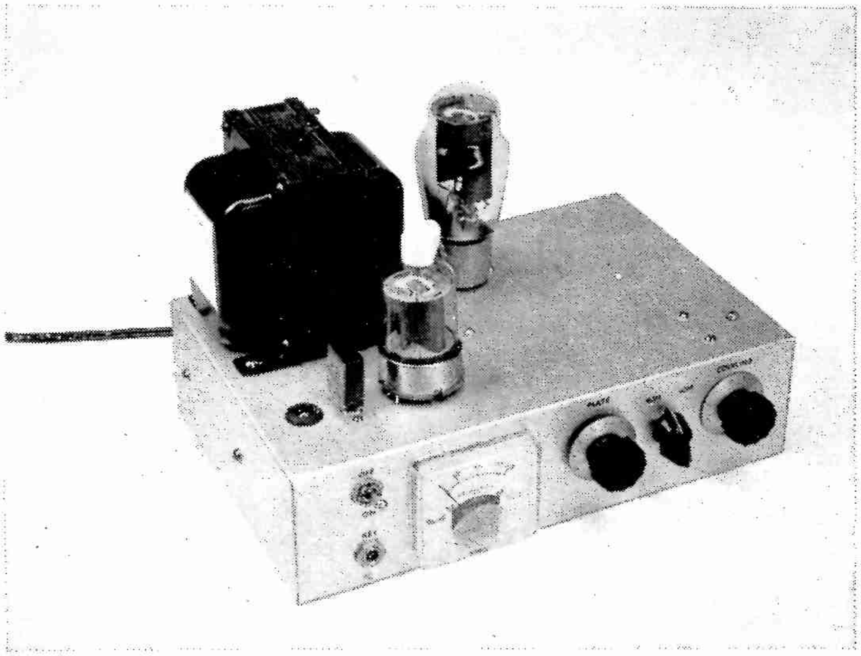
Construction

The transmitter is built on a 7 × 11 × 3-inch aluminum chassis. The meter requires a 2-inch hole, and the two tube sockets (Amphenol type MIP) take 1⅛-inch holes. The power transformer is mounted in the left rear corner of the chassis with the rectifier tube alongside. The crystal socket and 6146 tube are placed close together in front of the transformer. The lamp I_1 is mounted in a ½-inch rubber grommet set in the chassis close to the crystal socket. Connections to the lamp are made by soldering directly to its terminals.

On the front wall of the chassis, the power switch and key jack are mounted at the left-hand end. On the other side of the meter are the plate tank capacitor C_2, the band switch and the output capacitor C_4.

On the under side of the chassis, the filter choke is fastened against one end wall, and the

Fig. 6-39—This view of the 50-watter shows the panel arrangement and layout of the components above chassis. The crystal is between the 6146 and dial-light grommet. Behind the 6146 is the power transformer and to its right is the rectifier tube.



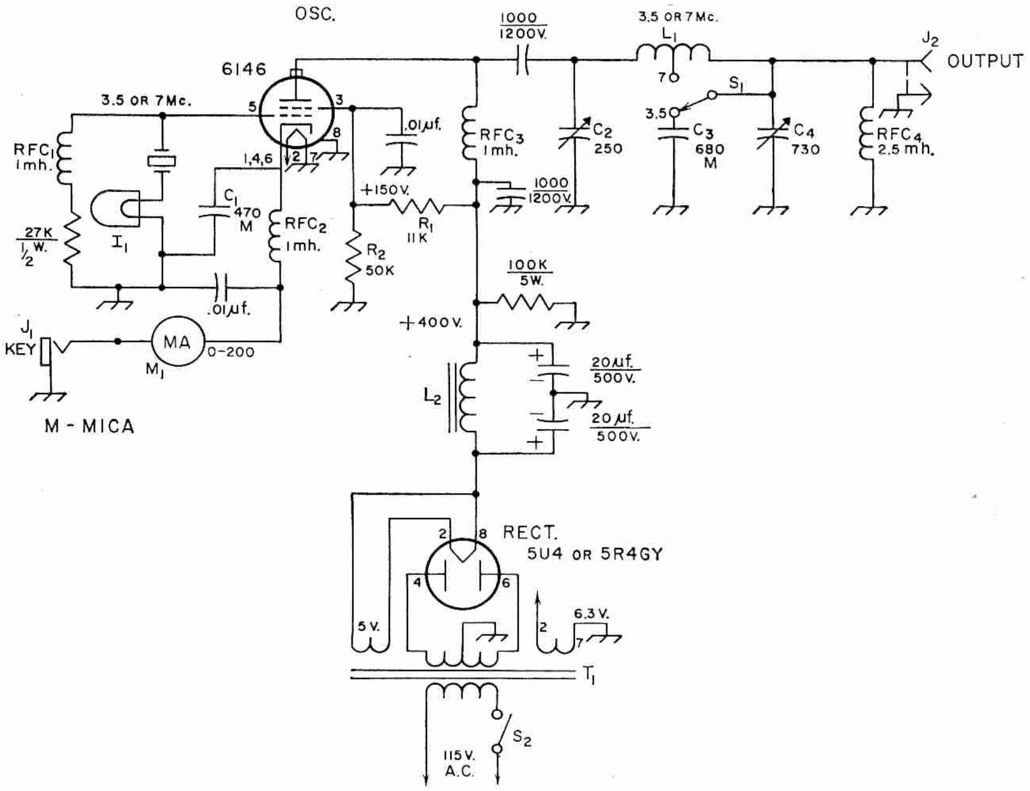


Fig. 6-40—Circuit diagram of the Novice-50 watter. Unless otherwise specified, capacitances are in µµf. Capacitors marked with polarity are electrolytic. Capacitors not otherwise identified are disk ceramic.

C_1—470-µµf. mica capacitor.
C_2—250-µµf. variable capacitor (Hammarlund MC-250M).
C_3—680-µµf. mica capacitor.
C_4—365-µµf.-per-section dual variable capacitor, broadcast-replacement type, sections connected in parallel (Allied Radio 60H725).
I_1—Dial lamp, 2 volts, 60 ma., No. 48 or 49.
J_1—Key jack, open-circuit.
J_2—RCA type phono jack.
L_1—35 turns No. 20, 1¼-inch diam., 16 t.p.i., tapped 15 turns from the C_4 end (B & W No. 3019).
L_2—9-hy. 125-ma. filter choke (Triad C-10X or equiv.).
M_1—2½-inch square (Shurite 850).
R_1—11,000 ohms 3 watts. (See text.)
R_2—50,000 ohms, 2 watts. (See text.)
RFC_1, RFC_2, RFC_3—1-mh. r.f. choke (National R-50, Millen 34300–1000).
RFC_4—2.5-mh. r.f. choke (National R-100S).
S_1—1-pole 2-position switch (Centralab No. 1460).
S_2—Single-pole single-throw toggle switch.
T_1—750 volts, c.t., 150 ma., 5 volts 3 amp., 6.3 volts, 4.5 amp. (Stancor PC-8411 or equiv.).

filter capacitors are against the rear wall, supported at the positive end by an insulated terminal strip, and at the negative end by soldering to the grounded terminal of the phono jack used as an output connector.

The coil L_1 is suspended by its leads between the stator terminals of the tank capacitor C_2 and the output or loading capacitor C_4.

On the 6146 socket, the three cathode prongs, Nos. 1, 4 and 6, should be connected together and the leads from C_1 and RFC_2 should be soldered to any of the three prongs.

On S_1, the center terminal connects to the stators of C_4. The 40-meter tap from L_1 goes to one outside terminal on S_1, and the mica capacitor C_3 goes to the other terminal.

Operation

After completing the wiring, check all connections to make sure you haven't made a mistake. When you feel you are ready to try the transmitter, plug in the key, an 80-meter crystal, the line cord, and turn the power on. Leave the key open until the 6146 warms up. A 40-watt light bulb makes a good load for testing the transmitter, the threaded portion connecting to the chassis ground and the base pin to the output lead.

Switch S_1 to the 80-meter position and set C_4 at maximum capacitance (plates fully meshed). Close the key and tune C_2 for a "dip" in meter reading. Once you've resonated the tank circuit by tuning C_2 to a dip, you may or may not find that the lamp lights. Also, the meter reading at the dip will probably be only 20 or 30 ma. By decreasing the capacitance of C_4 and redipping with C_2 you'll find that the lamp will get brighter and the loading heavier, as indicated by an increasing meter reading at the dip point. Be careful not to hold the key down any longer than necessary with the 6146 out of resonance as the tube is easily damaged during such operation. Increase the loading until the meter reads 100 to 125 ma. at the dip. This will be an input

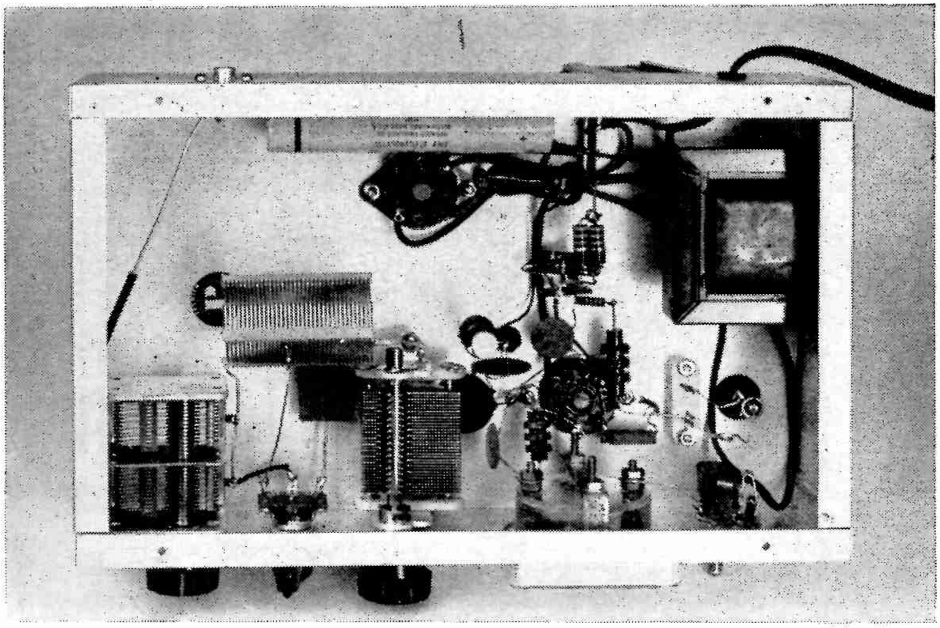


Fig. 6-41—This view shows the arrangement of the components below chassis. At the far right, mounted against the side of the chassis, is L_2, the power-supply choke. The filter capacitors are mounted along the back wall. At the lower left is C_4, the output capacitor. The other variable is C_2.

of approximately 50 watts, and the dummy load should be fairly bright. Under these conditions you should have approximately 400 volts on the plate of the 6146 and roughly 150 volts on the screen. Use an 80-meter crystal for 80-meter operation and a 40-meter one for 40. It is possible to use an 80-meter crystal for 40-meter work, but the oscillator will be operating as a frequency doubler and the output is less than when operating straight through at the crystal frequency.

Antennas

Antenna systems of any of the types discussed in the antenna chapter of this *Handbook* may be used with the transmitter, provided it is appropriate for the bands to be used. Two simple types of antenna are shown in the sketch of Fig. 6-42. Each will work on both of the two bands covered by the transmitter. The antenna shown in Fig. 6-42A consists of two dipoles, one for 80 meters and one for 40 meters, connected in parallel at the center where the feed line is attached. The antenna can be made of 300-ohm television ribbon line. First measure off two sections of ribbon each 66 ft. long. Then at the center of each section cut *one* of the two wires in the ribbon. Peel off *one* of the two 33-ft. sections of wire. Then connect the remaining 33-ft. wire and the 66-ft. section of the other conductor together as shown in the sketch. Repeat the same operation with the other 66-ft. section of ribbon line and attach an insulator between the two sections. The feed line should be connected across the insulator as shown.

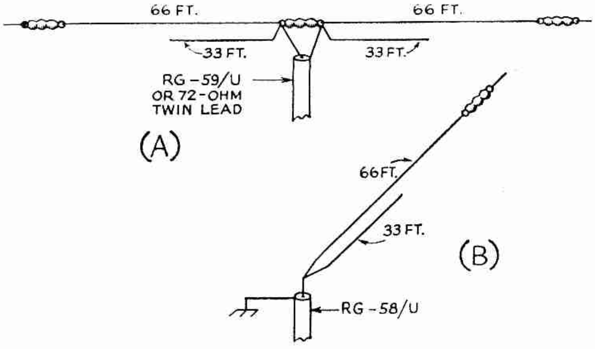


Fig. 6-42—Sketch of simple antennas described in the text. A shows a parallel-dipole system. The system of B requires a ground connection.

The antenna shown in Fig. 6-42B is similar in principle, except that the antennas are quarter-wave systems. This antenna is suitable if a good ground connection, such as a water pipe, is available within a few feet of the base of the antenna. The antenna is constructed in a manner similar to that described previously for the half-wave system. The antenna may be run vertically or run slanting to a tree or other support. If necessary, the first portion of it may be run vertically and the remainder horizontally.

The system of Fig. 6-42A should be fed with 72-ohm coax or ribbon line. The system of Fig. 6-42 should be fed with 52-ohm coaxial line.

To avoid possible second-harmonic radiation, particularly when operating in the 80-meter Novice band, an antenna tuner, such as the one described in *QST* for August, 1958, is recommended.

(Originally described in *QST* for December, 1958.)

A 75-Watt 6DQ5 Transmitter

The transmitter shown in Fig. 6-43 is designed to satisfy the requirements of either a Novice or General class licensee. As described here it is capable of running the full 75 watts limit in the 80-, 40- and 15-meter Novice bands, with bandswitching, crystal switching and other operating features. The General license holder can use the transmitter in any band 80 through 10 meters, and he can add v.f.o. control or amplitude modulation at any time without modifying the 6DQ5 transmitter. Crystal switching is a convenience for rapidly shifting frequency within a band to dodge QRM, and a SPOT position on the operate switch permits identifying one's frequency relative to others in a band. An accessory socket, X_3, furnishes a convenient point for borrowing power for a v.f.o. or for controlling the oscillator by an external switch.

Referring to Fig. 6-44, the circuit diagram of the transmitter, the crystal selector switch, S_1, is used to choose the desired crystal. For crystal-controlled operation crystals would be plugged in pins 1 and 3 and 5 and 7 of socket X_1. Similar sockets (not shown in the diagram) are used to hold the other crystals. When v.f.o. operation is desired, the v.f.o. output is connected to J_1, the plug P_1 is inserted in socket X_1, and the former 6AG7 crystal oscillator stage becomes an amplifier or multiplier stage when switch S_1 is turned to position 1.

Since the output of the 6AG7 stage will vary considerably with the bands in use, an excitation control, R_1, is included to allow for proper adjustment of the drive to the 6DQ5 amplifier. The 6DQ5, a highly sensitive tube, is neutralized to avoid oscillation; the small variable capacitor C_2 and the 390-$\mu\mu$f. mica capacitor form the neutralizing circuit. Screen or screen and plate modulation power can be introduced at socket X_2; for radiotelegraph operation these connections are completed by P_2. Grid or plate current of the 6DQ5 can be read by proper positioning of S_5; the 0–15 milliammeter reads 0–15 ma. in the grid-current position and 0–300 ma. in the plate-current position.

The transmitter is keyed at J_3, and a key-click filter (100-ohm resistor and C_5) is included to give substantially click-free keying. The v.f.o. jack, J_4, allows a v.f.o. to be keyed along with the transmitter for full break-in operation.

Construction

A 10 × 17 × 3-inch aluminum chassis is used as the base of the transmitter, with a standard 8¾-inch aluminum relay rack panel held in place by the bushings of the pilot light, excitation control and other components common to the chassis and panel. The panel was cut down to 17 inches in length so that the unit would take a minimum of room on the operating table. A good idea of the relative location of the parts can be obtained from the photographs. The support for the r.f. portion housing is made by fastening strips of 1-inch aluminum angle stock (Reynolds aluminum, available in many hardware stores) to the panel and to a sheet of aluminum 9½ inches long that is held to the rear chassis apron by screws and the key jack, J_3. A piece of aluminum angle must also be cut to mount on the chassis and hold the cane-metal (Reynolds aluminum) housing. Fig. 6-45 shows the three clearance holes for the screws that hold this latter angle to the chassis after the cane metal is in place. Build the can-metal housing as though the holes weren't there and the box has to hold water; this will minimize electrical leakage and the chances for TVI. To insure good electrical contact between panel and angle stock, remove the paint where necessary by heavy applications of varnish remover, with the rest of the panel

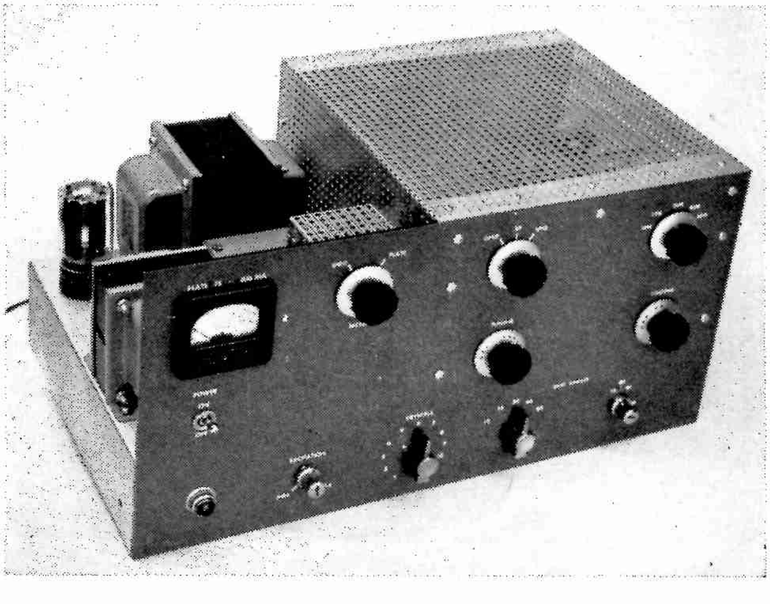


Fig. 6-43 — This 75-watt crystal-controlled transmitter has provision for the addition of v.f.o. control. A 6AG7 oscillator drives a 6DQ5 amplifier on 80 through 15 meters.

As a precaution against electrical shock, the meter switch, to the immediate right of the meter, is protected by a cane-metal housing. The switch to the right of the meter switch handles the spot-operate function, and the switch at the far top right is the plate-circuit band switch.

Along the bottom, from left to right: pilot light, excitation control, crystal switch, grid circuit band switch, and grid circuit tuning.

A 75-Watt Transmitter

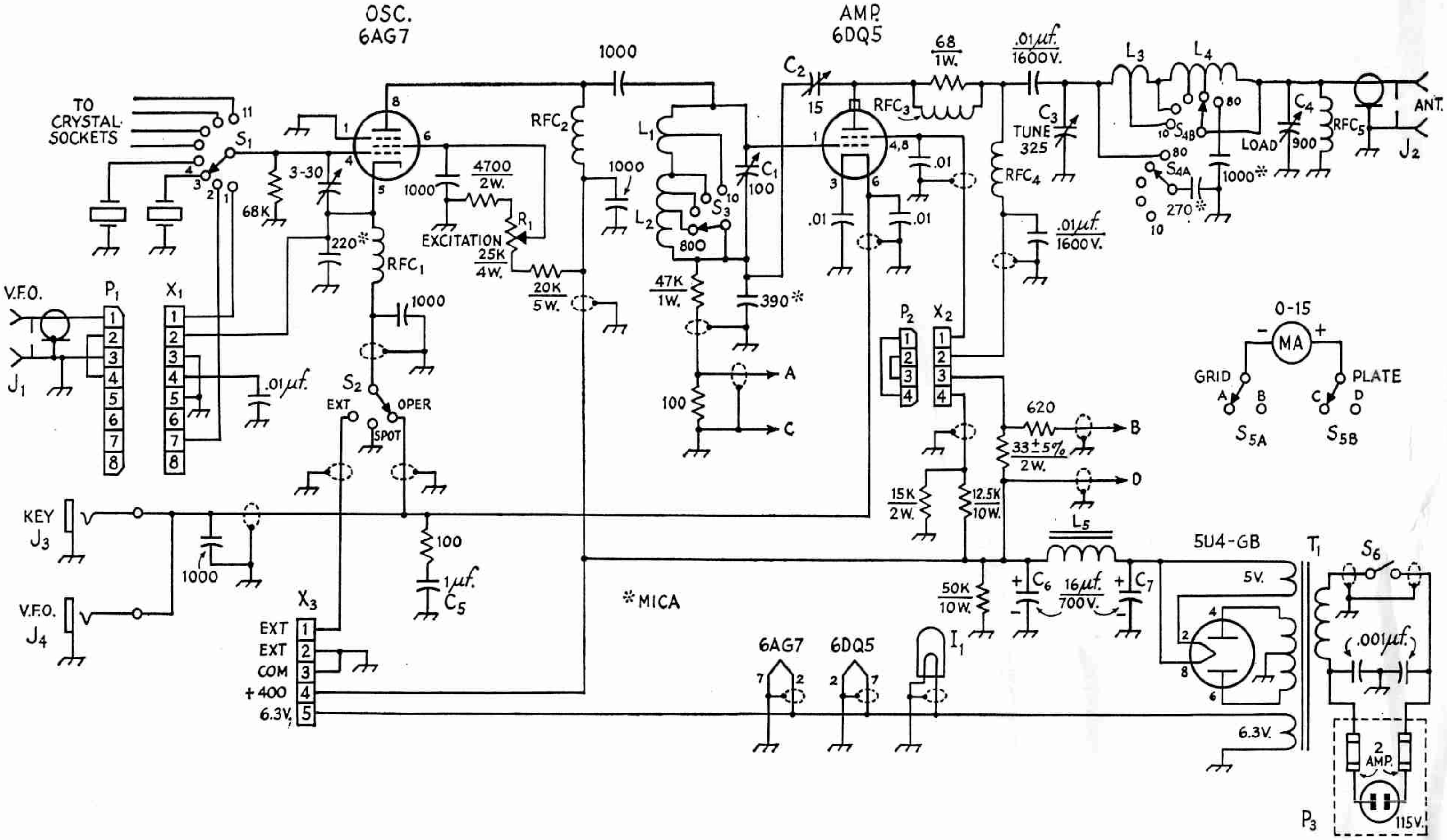


Fig. 6-44—Circuit diagram of the 75-watt 6DQ5 transmitter. Unless specified otherwise, capacitance is in µµf., resistance is in ohms, resistors are ½ watt.

C_1—100-µµf. midget variable (Hammarlund HF-100).

C_2—15-µµfd. midget variable, .025 inch spacing (Johnson 15J12).

C_3—325-µµf. variable (Hammarlund MC-325-M).

C_4—Dual 450-µµf. broadcast replacement variable, two sections connected in parallel.

C_5—1-µf. 400-volt tubular.

C_6, C_7—16-µf. 700-volt electrolytic (Aerovox PRS).

I_1—6-volt pilot lamp.

J_1—Phono jack.

J_2—Coaxial connector, chassis mounting, type SO-239.

J_3,J_4—Open-circuit phone jack.

L_1—7½ t. No. 18, ⅝ inch diam., 8 t.p.i., tapped 5½ turns from grid end (B&W 3006).

L_2—38 t. No. 32, 1 inch diam., 32 t.p.i., tapped 23 and 31 turns up (B&W 3016).

L_3—5 turns No. 14, 1-inch diam., 4 t.p.i., self-supporting, tapped 3½ turns from plate end.

L_4—15 turns No. 14, 1¾ inch diam., 4 t.p.i., tapped 6¼ and 10¼ from output end (B&W 3021).

L_5—10-henry 200-ma. filter choke (Triad C-16A).

P_1—Octal plug (Amphenol 86-PM8).

P_2—4-pin plug (Amphenol 86-PM4).

P_3—Fused line plug.

R_1—25,000-ohm 4-watt potentiometer (Mallory M25MPK).

RFC_1, RFC_2—750-µh. 100-ma. r.f. choke (National R-33).

RFC_3—3 turns No. 14 around 68-ohm 1-watt composition resistor.

RFC_4—1-mh. r.f. choke, 500 ma. (Johnson 102-752).

RFC_5—2.5-mh. r.f. choke (National R-100S).

S_1—1-pole 11-position rotary ceramic switch (Centralab Y section on P-121 index assembly).

S_2—Single-pole 11-position (3 used) non-shorting rotary switch (Centralab PA-1001).

S_3—Single-pole 12-position (5 used) rotary ceramic switch (Centralab PA-1 on PA-301 index assembly).

S_4 — 2-pole 5-position rotary ceramic switch (Centralab 2505).

S_5—S.p.s.t. toggle.

T_1—800 v.c.t. 200-ma. power transformer (Triad R-21A).

X_1—Octal tube socket.

X_2—4-pin tube socket.

X_3—5-pin tube socket.

masked off. The paint will blister and be easy to remove; wash the panel and then drill the holes for the components and screws. (If the holes are drilled first, the varnish remover may leak through and spoil the paint on the front of the panel.)

From a suitable piece of cane metal, make the four-sided 2¼ × 2¼ × 2¼-inch box that covers S_5, and fasten it to the utility-box cover with sheet-metal screws. Don't forget J_1 on the side of the box.

The self-supporting coil, L_4, can be wound on the envelope of the 6AG7 and then pulled apart to give the correct winding length.

Installation of the electrical components should present no problems. To insulate it from the chassis, capacitor C_1 is mounted on a small ceramic cone insulator (Johnson 135–500 or National GS-10). The socket for the 6DQ5 is mounted above the chassis on a pair of ¾-inch sleeves, with a large clearance hole under the socket for the several leads running from under the chassis. Cathode and screen bypass capacitors for the 6DQ5 connect to the chassis at soldering lugs under the sleeves.

Taps on L_2 are readily made by first pushing the wire on either side of the desired turn toward the center of the coil.

Note that shielded wire is used for many of the power leads; this is done to minimize the chances for stray radiation and it also contributes to the stability of the transmitter. Don't neglect it.

Adjustment

When the wiring is completed and checked, disable the amplifier stage by removing P_2, plug in P_3 and turn on S_5. The tube heaters and filaments should light up. If a voltmeter is available and connected across C_6, it should indicate over 500 volts. Later on, with full loading, the plate voltage will run around 400.

With S_1 switched to an 80-meter crystal, S_3 switched to 80 or 40 and S_5 switched to GRID, flip S_2 to SPOT and tune C_1 through its range. If the crystal is oscillating the meter should give an indication at some setting of C_1. The grid current reading should vary with the setting of C_1 (maximum at resonance) and with the setting of R_1 (maximum with arm at 20K end). If a key is plugged in at J_3 and S_2 is set to OPER, the grid current should appear only when the key is closed. Listen to the signal on a receiver (no antenna); if the signal is chirpy try adjusting the 3–30 μμf. compression trimmer between grid and cathode of the 6AG7.

With a 40-meter crystal switched in, check for grid current at 14 and 21 Mc., by switching S_3 to the desired band and tuning with C_1. These settings should be checked with an absorption-type wavemeter, since it is possible in some cases to find more than one harmonic in the range of C_1. The 28-Mc. range can also be checked, but the 4th harmonic of the 7-Mc. crystal will yield only about 1 ma. of grid current.

Next check the neutralization on the 15-meter

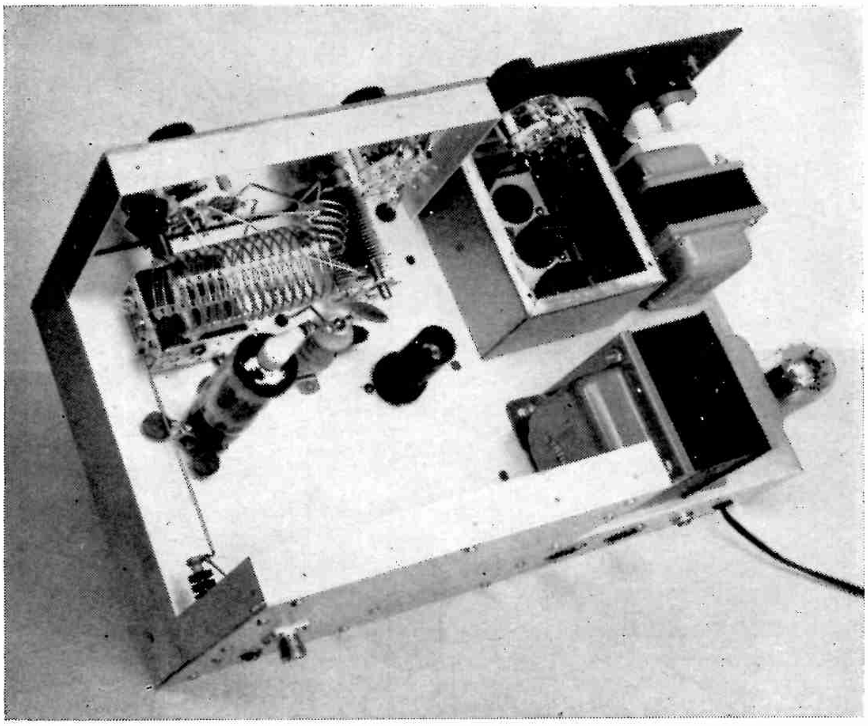


Fig. 6-45 — Top view of the 6DQ5 transmitter with cane-metal cover removed. A 3 × 4 × 5-inch utility box (upper right) serves as a shield for the crystals; the cane-metal protection for the meter switch is fastened to the box cover. Phono jack mounted on the meter-side of the box receives v.f.o. output; short length of Twin-Lead from this jack to octal plug brings v.f.o. output to crystal socket.

For protection against high voltage, meter terminals are covered by ceramic tube plate caps (Millen 36001).

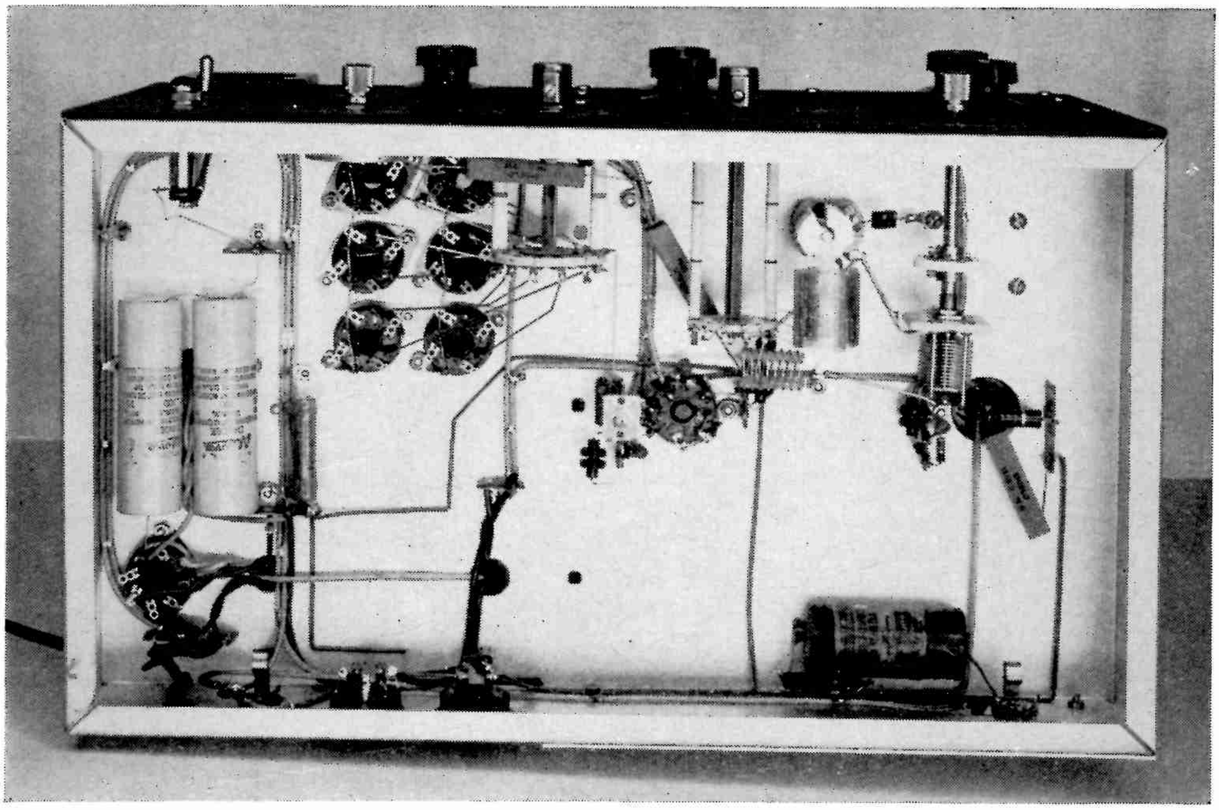


Fig. 6-46 — Group of six octal sockets (upper left) serves as crystal sockets. Socket at center of chassis holds 6AG7 oscillator tube; the 3–30-$\mu\mu$f. mica compression trimmer mounted alongside is excitation control for oscillator stage. Small midget capacitor above coil is neutralizing capacitor adjusted from above chassis; this capacitor and grid tuning capacitor to right must be insulated from chassis.

band. With 21-Mc. grid current indicating, switch S_4 to 15, set C_4 at half scale, and swing C_3 through its range. Watch closely for a flicker in grid current. If one is observed, try a different setting of C_2. Work carefully until the flicker is a minimum. A more sensitive indication of neutralization can be obtained by using a germanium diode and a 0–1 milliammeter in the output at J_2; adjust C_2 for minimum meter indication. If using this sensitive test, it is wise to start out with R_1 set at half range or less, until it has been determined that the meter will not swing off scale. Under no circumstances use this test with P_2 in place; the 6DQ5 output is quite likely to destroy the crystal diode.

When the amplifier has been neutralized, connect a dummy load (a 60-watt lamp will do) at J_2 and replace P_2. Set S_5 to PLATE and send a few dots as C_3 is tuned through its range. At resonance the lamp should light up and the plate current should dip. The plate current can be made to increase, along with the lamp brilliance, by decreasing the capacitance at C_4. The 6DQ5 plate current can be run up to 180 ma. (9 ma. on the meter) for Novice work; the grid current should be held at 2 to 4 ma. Crystals in the 3.5- to 4.0-Mc. range should be used for 80- and 40-meter operation, and 7-Mc. crystals should be used on 40, 20 and 15 meters. For 10-meter operation, it is recommended that a v.f.o. with 20-meter output be used to drive the 6AG7; trying to drive the 6DQ5 with the 4th harmonic of a 7-Mc. crystal is too marginal for all but the most experienced operators. With v.f.o. control, always frequency multiply (double or triple) in the 6AG7 stage to the desired band.

Because the 6DQ5 is capable of drawing high values of plate current when not tuned properly, it will pay to take care in learning how to adjust the transmitter. Once the controls have been "calibrated" and the approximate settings for each band become known, it should no longer be necessary to tune up with the "series-of-dots" technique mentioned above. However, in the early stages of familiarization with the transmitter, the dots, or a fast hand on the key, may save a tube or power supply. The fact that the 6DQ5 can draw such heavy currents at low plate voltages makes it an excellent tube for an effective inexpensive transmitter, but the tube is not as tolerant of careless tuning habits as are some other tubes.

A 90-Watt All-Purpose Amplifier

The amplifier shown in Figs. 6-47 through 6-50 will serve as a Class-AB_1 linear amplifier or as a Class-C power amplifier with no changes other than the proper adjustment of excitation and loading. To accomplish this, a stabilized bias supply provides proper Class-AB_1 bias; the bias increases to the correct value for Class-C operation when the excitation is brought up to the point that yields normal grid current. A stabilized screen supply is included to insure good linear operation.

Referring to the amplifier circuit in Fig. 6-49, excitation on the desired band is introduced at J_1. The grid circuit is a commercial assembly, Z_1, that can be switched to the correct band by S_1 and tuned by C_1. A pi-network coupler is used in the output, switched by S_2 and tuned by C_3. Proper loading is obtained by adjustment of C_4; to provide sufficient output capacitance in the 80-meter band an additional 680 $\mu\mu$f. is added. A neutralizing circuit, C_2 and a 680-$\mu\mu$f. capacitor, adds to the fundamental stability at the higher frequencies. Parasitic suppressors were found to be necessary in the grid and plate circuits.

Overload protection is provided by a 250-ma. fuse in the cathode circuit. The grid, plate or screen current can be metered by a suitable setting of S_3; with the resistances shown the meter provides a full-scale reading of 5 ma. on grid current, 25 ma. on screen current, and 250 ma. on plate current.

If it is desired to plate- or screen-modulate the amplifier for a.m. operation, the necessary audio power can be introduced at J_3.

Fig. 6-47—Front view of the 6146 all-purpose amplifier The upper panel is part of an 8 × 6 × 3½-inch Minibox (Bud CU-2109); the ventilated shielding of Reynolds Aluminum cane metal is fastened to the Minibox and base with sheet-metal screws.

Plate-circuit tuning controls and switch are mounted on the Minibox, and the grid-circuit controls, power switches and meter are mounted on the end of the 8 × 12 × 3-inch aluminum chassis that serves as a base.

The power-supply circuit is shown separately (Fig. 6-51) for convenience only, since the amplifier and power supply are all built on the same 8 × 12 × 3-inch chassis. High voltage for the plate of the 6146 is provided by a bridge rectifier using a 5U4-GB and two 6DE4 rectifiers; stabilized screen voltage is obtained from the same supply and two voltage-regulator tubes.

Fig. 6-48—Rear view of the 90-watt all-purpose amplifier with the cane-metal cover removed. One voltage-regulator tube has been removed from its socket (right edge of transformer) to allow the neutralizing capacitor and plate blocking capacitor to be seen. The plate r.f. choke (RFC_3 in Fig. 6-49) is mounted on one side wall, and the load capacitor and safety choke (C_4 and RFC_4 in Fig. 6-49) are mounted on the far side wall.

The rear apron of the chassis (foreground) carries the input and output coaxial-connector jacks, the 6146 cathode fuse, and the socket for the a.m. modulator connections. A shorting plug is shown in the socket.

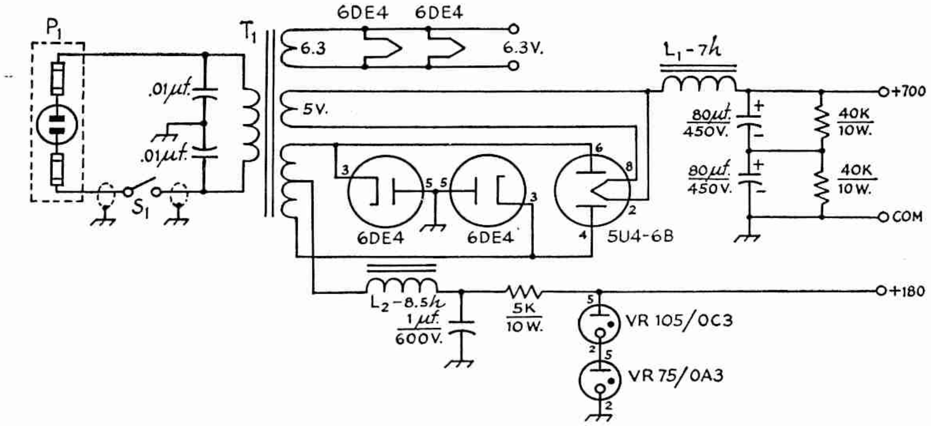


Fig. 6-51—Power supply section of the all-purpose amplifier.

L_1—7-henry 150-ma. choke (Stancor C-1710).
L_2—8½-henry 50-ma. choke (Stancor C-1279).
P_1—Fused line plug, 3-ampere fuses.
S_1—S.p.s.t. toggle.
T_1—800 v.c.t. at 200 ma., 6.3 v. at 5 amp., 5 v. at 3 amp. (Allied Radio Knight 62 G 033).

tion source to be used — less than a watt is required for linear operation, and only a shade more for Class C. Use the drive at a high frequency, such as 21 or 28 Mc. Turn on the amplifier and switch the band switches to the band corresponding to the excitation-source frequency. Adjust the grid tuning capacitor for a show of grid current; peak the tuning and (if necessary) adjust the excitation for a half-scale reading of grid current. With the loading capacitor C_4 set at half scale, swing the tuning capacitor C_3 through its range. Watch carefully for a slight flicker in grid current. If one is found, adjust the neutralizing capacitor C_2 until the flicker is minimized. The amplifier is now neutralized. Alternatively, a sensitive detector of r.f. can be coupled at the output connector, J_2, and used instead of the grid-current flicker. Adjust C_2 for minimum r.f. in the output when the plate circuit is tuned through resonance. Turn off the power switch and disconnect the excitation source.

Remove the sensitive detector, if used, and replace the rectifier tubes. Turn on the power and switch the meter to read plate current. With the grid and plate circuits switched to the same band (10, 15, 20 or 40) it should be possible to swing the grid and plate tuning to any combination of settings with no change in plate current reading. This indicates that the amplifier is stable and free from oscillation. (The amplifier can be made to oscillate on 80 meters with no grid or plate loading, but in loaded operation it will be stable.)

The antenna and excitation can now be connected and the amplifier used in normal fashion. Used as a linear amplifier, the excitation should be adjusted just below the level that would kick the grid-current indication on signal peaks. Proper loading will be obtained when a steady carrier just under the grid-current level is used for drive and the loading at resonance is set for about 100 ma. plate current. Under these conditions of loading, a sideband signal will kick the plate current to about 40 or 50 ma. on peaks. Measured p.e.p. input before clipping should be 60 to 70 watts.

When used as a Class-C amplifier, the drive should be increased to where about 2 to 3 ma. grid current is drawn, and the loading to where the 6146 draws about 125 ma. If the amplifier is plate modulated, the plate current should be reduced to 95 ma., to stay within the tube ratings.

Since the amplifier uses a fixed and "stiff" screen supply, it is good practice always to bring up the excitation and loading together, while checking to see that the screen current never exceeds about 15 ma. In normal Class-C operation the screen current will run around 10 ma.

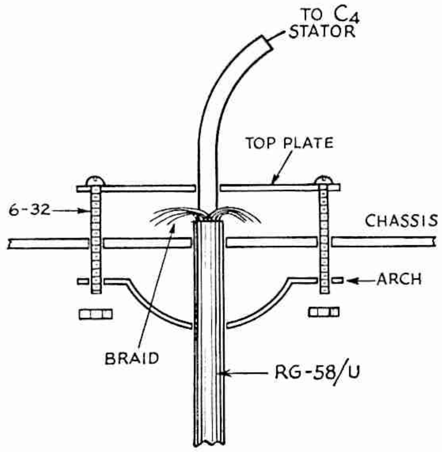


Fig. 6-52—Exploded view of the cable clamp used to hold the coaxial cable running to J_2. The top plate is a 1½-inch square of sheet aluminum with holes at the four corners for 6-32 screws. The arch is a 7/16-inch wide strap that mounts diagonally under the chassis. When tightened, the top plate clamps the cable braid to the chassis; the arch lends support to the cable.

6—HIGH-FREQUENCY TRANSMITTERS

A Self-Contained 500-Watt Transmitter

Figs. 6-53 through 6-58 show the details of a 500-watt c.w. transmitter, completely self-contained except for the external remote v.f.o. tuning box shown in Figs. 6-57 and 6-58. Provision is made for introducing s.s.b. input at the grid of the driver stage. While plate modulation can be applied to the final amplifier in the usual manner, ratings of the plate power supply limit the safe input to about 250 watts.

The circuit is shown in Fig. 6-54. Switch S_2 permits either v.f.o. or crystal-controlled operation using a 6AH6 oscillator. Either 80- or 40-meter crystals may be used. The v.f.o. circuit is in the 80-meter band and S_1 selects either of two frequency ranges — 3.5 to 4 Mc. for complete coverage of all bands, and 3.5 to 3.6 Mc. for greater bandspread over the low-frequency ends of the wider bands. The plate circuit of the oscillator is on 40 meters for all output bands except 80 meters where it is non-resonant.

A 6CL6 buffer separates the oscillator and the first keyed stage. This stage doubles to 20 meters for 20- and 10-meter output and triples to 15 meters. The driver is a 2E26 which doubles to 10 meters and works straight through on all other bands. This stage is neutralized and a potentiometer in its screen circuit serves as an excitation control.

The final is a 7094, also neutralized, with a pi-network output circuit using a B&W 851 bandswitching inductor unit.

A differential break-in keying system using a 12AU7 is included. Both the final amplifier and driver are keyed by the grid-block method. The differential is adjusted by R_1. Clicks are prevented by envelope-shaping circuits which include C_7, C_{11} and the grid-leak resistances.

The 100-ohm meter shunts give a full-scale reading of 50 ma., the 51-ohm shunts a full-scale reading of 100 ma., and the 10-ohm resistor in the negative high-voltage lead provides a 500-ma. scale.

Power Supply

The plate transformer in the high-voltage supply uses a transformer designed for a conventional full-wave rectifier circuit with an ICAS d.c. output rating of 300 ma. at 750 volts. A bridge rectifier is used with this transformer so that an output voltage of 1500 is obtained. The short duty cycle of c.w. or s.s.b. operation makes it possible to draw up to the rated maximum of the 7094 (330 ma.) through a choke-input filter without a prohibitive rise in transformer temperature.

The low-voltage supply has two rectifiers. A full-wave rectifier with a capacitive-input filter provides 400 volts for the plate of the driver and the screen of the final amplifier. A tap on a voltage divider across 400 volts provides 300 volts for the plates of the oscillator, buffer and keyer tubes. A half-wave rectifier with a choke-input filter supplies 250 volts of bias for the keyer and fixed bias for the 2E26 and 7094 when they are operating as Class AB_1 linear amplifiers.

Control Circuits

S_7 is the main power switch. It turns on the low-voltage, filament and bias supplies. Until it has been closed, the high-voltage supply cannot be turned on. In addition to turning on the high-voltage supply, S_8 operates the relay K_1 which applies screen voltage to the final amplifier. Thus, to protect the screen, screen voltage cannot be applied without applying plate voltage simultaneously. J_8 is in parallel with S_8 so that the high-voltage supply can be controlled remotely from an external switch. Also, in parallel with the primary of the high-voltage transformer is another jack, J_7, which permits control of an antenna relay or other device by S_8 if desired.

The v.f.o.-set switch S_5 turns on the exciter and grounds the screen of the final amplifier.

S_2 has three positions. One is for crystal control, the second for v.f.o. operation, and the third position is for operating the last two stages of the transmitter as linear amplifiers with an external s.s.b. exciter. In addition to shifting the input of the driver stage from the buffer amplifier to an s.s.b. input connector, fixed bias is provided for AB_1 operation of both stages.

Construction

The transmitter is assembled on a 17 × 13 ×

Fig. 6-53—A 500-watt transmitter. Power supplies and a differential keyer are included. It operates with the external v.f.o. tuner shown in Fig. 6-57. Controls along the bottom, from left to right, are for low-voltage power, v.f.o./crystals/s.s.b. switch, driver tank switch, driver tank capacitor, final loading, v.f.o. set switch, and high-voltage. Above, from left to right, are controls for excitation, final tank switch, final tank capacitor and meter switch. The band-switch pointer is made by cutting down the metal skirt of a dial similar to the one to the right. All dials are Johnson.

500-Watt Transmitter

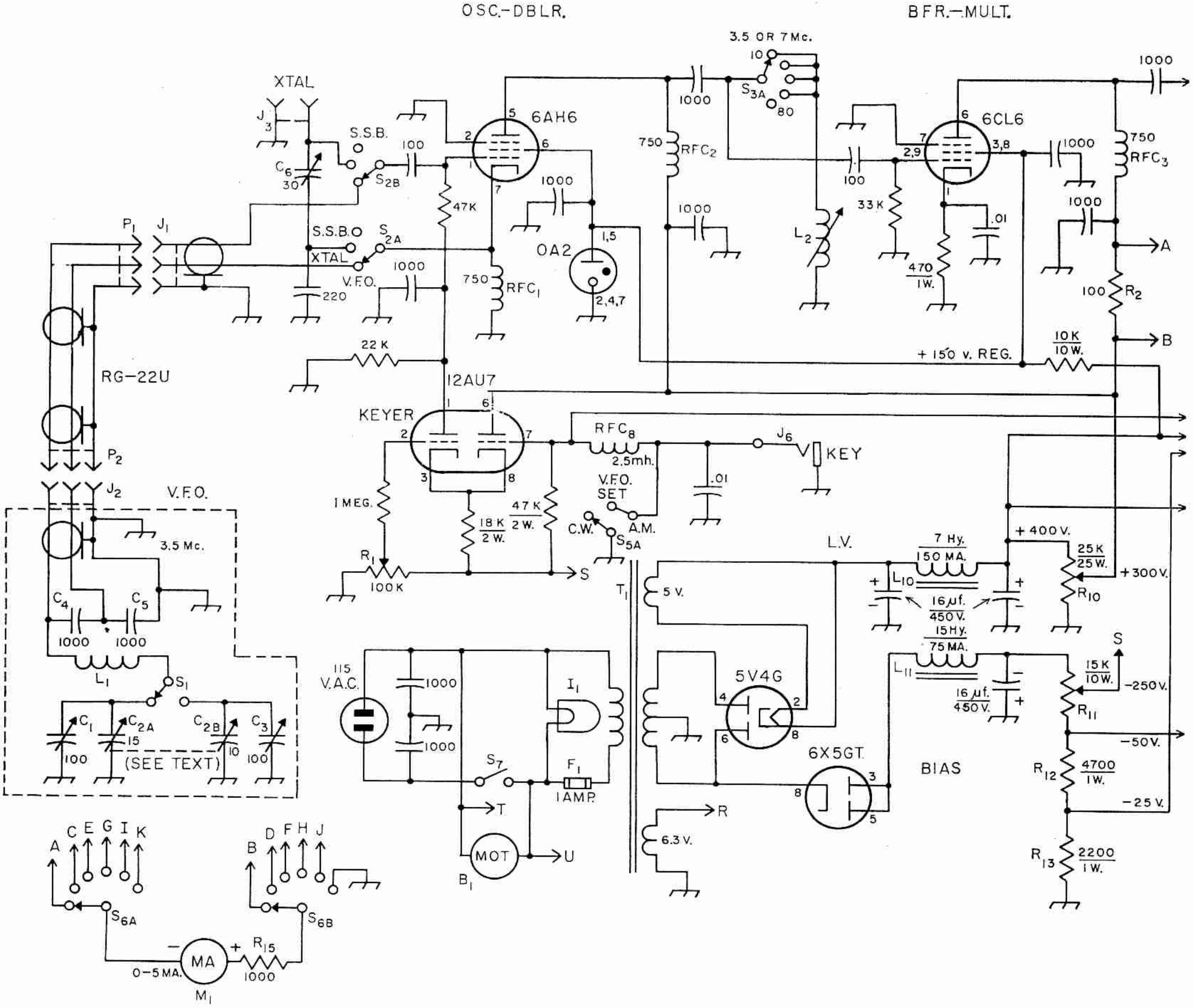


Fig. 6-54—Circuit of the 500-watt self-contained transmitter. Capacitance less than 0.001 μf. are in μμf. Fixed capacitors of capacitance greater than 100 μμf. should be disk ceramic, except as noted below. Fixed capacitors of 100 μμf. and 220 μμf. should be mica. Capacitors marked with polarity are electrolytic. Resistors not otherwise marked are ½ watt.

B_1—Blower (Allied 72P715).

C_1, C_3—100-μμf. air trimmer (Hammarlund APC-100-B).

C_2—Midget dual variable, 25 μμf. per section (Johnson 167-51 altered as described in the text).

C_4, C_5—0.001-μf. silver mica.

C_6—30-μμf. mica trimmer (National M-30).

C_7, C_{11}—0.1-μf. paper (keyer shaping).

C_8—30-μμf. miniature variable (Johnson 160-130).

C_9—100-μμf. midget variable (Johnson 167-11).

C_{10}—330-μμf. mica.

C_{12}—10-μμf. neutralizing capacitor (Johnson 159-125).

C_{13}—0.001-μf. 3000-volt disk ceramic.

C_{14}—0.001-μf. 5000-volt ceramic (CRL 858S).

C_{15}—250-μμf. 2000-volt variable (Johnson 154-1).

C_{16}—Triple-gang broadcast variable, 365 μμf. or more per section, sections connected in parallel.

I_1, I_2—One-inch 115-volt panel lamp.

J_1, J_2—Cable connector for RG-22/U (Amphenol 83-22R, UG-103/U).

J_3—Crystal socket (Millen 33102).

J_4, J_5—Coaxial receptacle (SO-239).

J_6—Key jack, open circuit.

J_7, J_8—Chassis-mounting a.c. receptacle (Amphenol 61-F).

K_1—S.p.s.t. 115-volt a.c. relay (Advance GHA/1C/115VA or similar).

L_1—35 μh.—32 turns No. 18, 2 inches diameter, 2 inches long (Airdux 1616).

L_2—Approx. 10 μh.—65 turns No. 26 enam., on ⅜-inch iron-slug form (Waters CSA-1011-3).

L_3—Approx. 2 μh.—16 turns No. 26 enam., close-wound at center of form similar to L_2.

L_4—Approx. 1 μh.—13 turns No. 26 enam., ½ inch long at center of form similar to L_2.

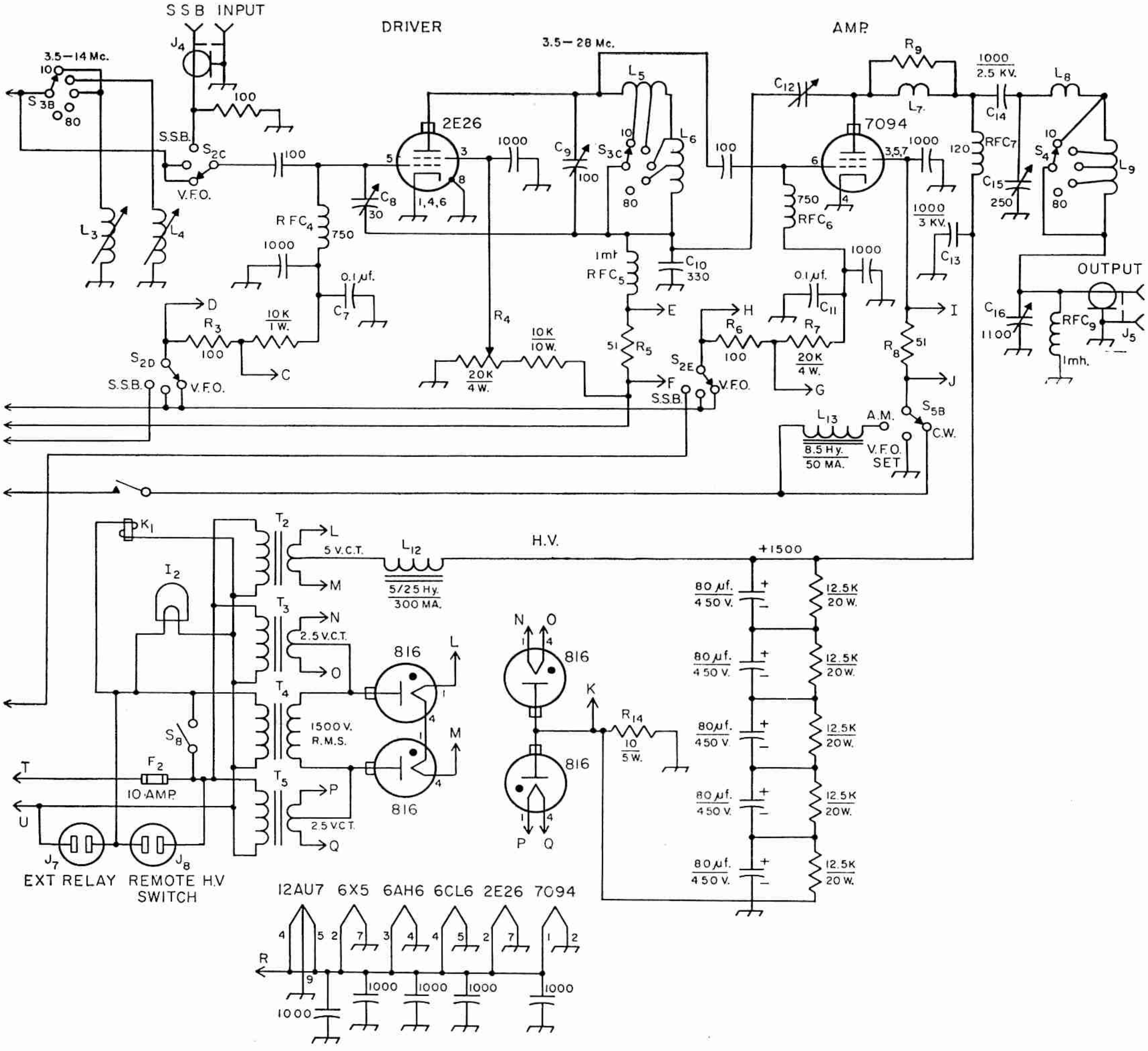
L_5—16 turns No. 20, ¾ inch diameter, 1 inch long, tapped at 10 turns and 13 turns from L_6 end (Airdux 616).

L_6—40 turns No. 16, 1¼ inches diameter, 2¾ inches long, tapped at mid point and at L_5 end (Airdux 1016).

L_7—3 turns No. 14, ½ inch diameter, ¾ inch long.

L_8—4 turns 3/16 × 1/16-inch copper strip, 1⅜ inches diameter, 2½ inches long (part of B&W 851 coil unit).

L_9—4¾ turns No. 8, 2½ inches diameter, 1¾ inches long, tapped at 1¾ turns from L_8 end, plus 9½ turns No. 12, 2½ inches diameter, 1½ inches long, tapped at 6 turns from output end (part of B&W 851 coil unit).



L_{10}—7-hy. 150-ma. filter choke (Stancor C-1710).

L_{11}—15-hy. 75-ma. filter choke (Stancor C-1002).

L_{12}—5/25-hy. 300-ma. swinging filter choke (Triad C-33A).

M_1—Shielded 0- 5-ma. d.c. milliammeter, 3½-inch rectangular (Phaostron).

P_1, P_2—Plug for RG-22/U cable (Amphenol 83-22SP).

R_1—100,000-ohm potentiometer.

R_2, R_3, R_6—100 ohms, 5%.

R_4—20,000-ohm 4-watt potentiometer (Mallory M20-MPK)

R_5, R_8—51 ohms, 1 watt, 5%.

R_7—Two 10,000-ohm 2-watt resistors in series.

R_9—Three 100-ohm 1-watt noninductive resistors in parallel.

R_{10}—25,000 ohms, 25 watts with slider.

R_{11}—15,000 ohms, 20 watts, with slider.

R_{12}—4700 ohms, 1 watt.

R_{13}—2200 ohms, 1 watt.

R_{14}—10 ohms (Five 51-ohm 1-watt 5% resistors in parallel.)

R_{15}—1000 ohms ½ watt 5%.

S_1—Single-pole ceramic rotary switch (Centralab 2000, 2 of 12 positions used).

S_2—Two-wafer ceramic rotary switch (Centralab PA-300 index, PA-4 wafers. S_{2A} and S_{2B} are on one wafer, S_{2C}, S_{2D} and S_{2E} on second wafer).

S_3—Three-wafer ceramic rotary switch (Centralab PA-301 index, wafers PA-0, 5 positions used).

S_4—Part of B&W 851 coil unit.

S_5—2-pole 3-position ceramic rotary switch (Centralab 2003, two positions used).

S_6—Double-pole ceramic rotary switch (Centralab 2003).

S_7, S_8—S.p.s.t. toggle switch.

T_1—Power transformer: 750 v.a.c., c.t., 150 ma.; 5 volts 3 amps.; 6.3 volts, 4.7 amps. (Thordarson 22R06).

T_2, T_3—Filament transformer: 2.5 volts, c.t., 3 amps. (Triad F-1X).

T_4—Plate transformer: 1780 volts, c.t., 310 ma., center tap not used (Triad P-14A).

T_5—Filament transformer: 5 volts, c.t., 3 amps. (Triad F-7X).

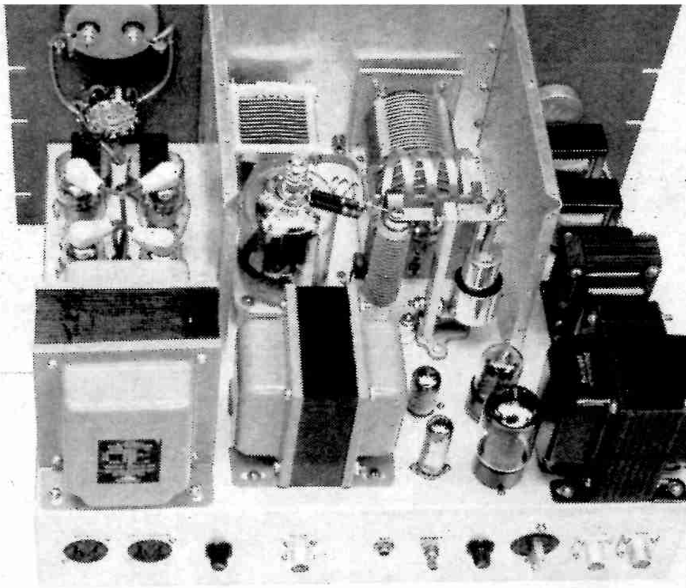


Fig. 6-55—The only shielding required on top of the chassis is the amplifier enclosure shown. A perforated cover for the enclosure is not shown.

4-inch aluminum chassis with a 19 × 12¼-inch panel. The amplifier enclosure measures 8½ inches wide, 8¼ inches deep and 7½ inches high. The three permanent sides shown in Fig. 6-55 can be bent up from a single sheet of solid aluminum stock. The top and back (not shown) are made from a single piece of Reynolds perforated sheet aluminum.

The tube socket is mounted on ¾-inch ceramic cones over a large hole cut in the chassis and covered with a patch of perforated sheet. The tank capacitor C_{15} is mounted on metal spacers to bring its shaft level up to that of the switch on the B&W inductor which is mounted directly on the chassis. The two shafts are spaced 4 inches.

Exciter

A 4 × 5 × 6-inch aluminum box is used as the foundation for the exciter. The driver tank capacitor is centered on the chassis with its center approximately 3 inches back from the front edge of the chassis. The capacitor specified has an insulated mounting. If an uninsulated capacitor is substituted, an insulating mounting must be provided. The shafts of S_2 and S_3 are spaced 2½ inches and centered on the front end of the box. On the side of the box toward the tuning capacitor, the oscillator tube, the buffer tube, the low-frequency section (L_6) of the driver tank coil, and the 2E26 are lined up so as to clear the tank capacitor and its shaft. The latter is fitted with an insulated coupling and a panel-bearing unit. The slug-tuned coils are mounted in holes near the bottom edge of the box. Neutralizing capacitor C_8 is mounted at the rear end of the box, close to the 2E26 socket. The high-frequency section (L_5) of the tank coil is suspended between the outer end of the low-frequency section and the plate cap of the 2E26. Coil-tap leads run through small feed-through points or grommeted clearance holes in the side of the box.

The loading capacitor C_{16} is placed so that its shaft is symmetrical with the shaft of S_3, and S_5 is spaced from it to balance S_2 at the other end.

The V.F.O. Tuner

The v.f.o. tuner is assembled in a 5 × 6 × 9-inch aluminum box (Premier AC-596). The dual tuning capacitor C_2 has 7 plates, 4 rotor and 3 stationary, in each section. In the front section, which is used to cover the entire 80-meter band, the two rotor plates nearest the front should be removed. This leaves two rotor plates and two active stator plates, the front stator plate being inactive. In the rear section, the front rotor plate and the last two rotor plates are removed. This leaves one rotor plate riding between two stators.

The capacitor is mounted on a bracket fastened against the bottom of the box, although it could be mounted from the front cover with spacers to clear the hub of the Millen 10035 dial. The shaft of the capacitor should be central on the front cover. The coil is suspended between a pair of

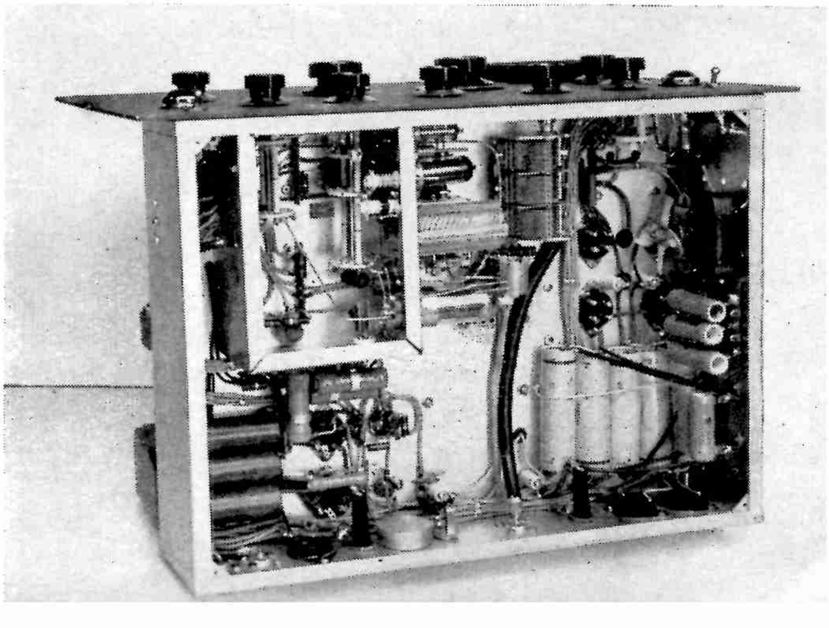


Fig. 6-56—The exciter is assembled using a standard aluminum box as the foundation. The perforated cover has been removed. The bottom of the chassis should also have a perforated metal cover.

$2\frac{1}{2}$-inch ceramic pillars (Millen 31002). It is placed immediately to the rear of the tuning capacitor. The two air trimmers, C_1 and C_3, are mounted on the top side of the box with their shafts protruding so that they can be adjusted from the top. The bandspread switch is mounted in one end of the box and the cable connector at the other end.

The unit is housed in a standard cabinet (Bud C-1781) having an 8 × 10-inch panel. The dial should be fastened to the panel, making sure that the hub of the dial lines up accurately with the shaft of the tuning capacitor. Then the box is inserted in the cabinet through the front opening. The switch shaft goes out through a hole drilled in the side of the cabinet, and the cable goes through a hole in the opposite end to the cable connector. The dial should be set to read zero at maximum capacitance of the tuning capacitor. The box should be supported on spacers.

Fig. 6-57—The remote v.f.o. tuning unit is housed in a standard metal cabinet. The cable at the right plugs into the main chassis.

Adjustment

With all tubes except the rectifiers out of their sockets, the power supplies should be checked first to be sure that they are functioning properly. The voltage output of the low-voltage supply should be in excess of 400 volts, the biasing voltage 300 or more and the high voltage above 1500. The slider on the low-voltage bleeder should be set at approximately three quarters of the way from ground. The slider on the bias-supply bleeder should be set for a reading of −250 volts to ground.

Plug in the oscillator and buffer tubes and an 80-meter crystal if one is available; otherwise connect the v.f.o. tuner. With the low-voltage supply turned on, the 0A2 should glow. When the key is closed, the 0A2 should dim but stay ignited. If it does not, the value of the 10K VR resistor should be reduced.

The v.f.o. can now be adjusted to frequency. Set C_2 at maximum capacitance. Set S_1 to the 80-meter position. Adjust the 80-meter trimmer until a signal is heard at 3500 kc. on a calibrated receiver. Then set the receiver to 4000 kc. and tune the v.f.o. until the signal is heard. If the signal is not close to 100 on the dial, carefully bend the rear rotor plate of the 80-meter section of C_2 outward a little at a time to get the desired bandspread. Each time this adjustment is made, the trimmer should be reset to bring 3500 kc. at zero on the dial.

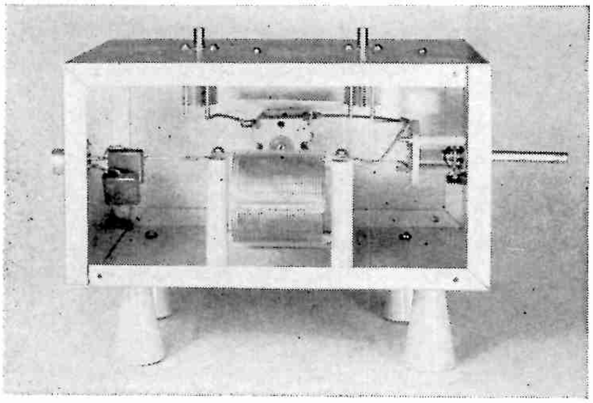


Fig. 6-58—Interior of the v.f.o. tuning box showing the mounting of the coil and other components.

The same procedure should be followed in adjusting for the other v.f.o. range, aiming for 3600 kc. (or above if desired) at 100 on the dial.

The 2E26 should now be plugged in and the excitation control R_4 set at the ground end (zero screen voltage). S_2 should be set in the v.f.o. position. With low voltage on and the key closed, a 5763 grid-current reading should be obtained with the band switch in the 80-meter position. With the switch in the 40-meter position, the slug of L_2 should be adjusted for maximum grid current to the 2E26. With the band switch in the 20-meter position, L_3 should be adjusted for maximum grid current, and then the slug of L_4 should be adjusted for maximum grid current with the band switch in the 15-meter position.

Now insert the 7094 in its socket and neutralize the 2E26 as described earlier in this chapter.

Testing of the final amplifier requires a load applied to the output connector. Two 150-watt lamps connected in parallel should serve the purpose. Turning on the high voltage will also apply screen voltage through the relay K_1. With both band switches set to 10 meters, and C_{16} set at about half capacitance, quickly tune the output circuit to resonance as indicated by the plate-current dip. The load lamp should show an indication of output. Switch the meter to read grid current and neutralize as described earlier in this chapter. After neutralization the amplifier can be loaded to rated plate current. If it is above the rated maximum value, increase C_{16} and retune to resonance, or decrease C_{16} if the plate current is below the rated value.

With the final adjusted and the entire transmitter operating, make a final check on the voltage at the tap on the low-voltage supply, adjusting the slider if necessary to bring the voltage to 300 with the key closed. Be sure to turn off *all* voltages each time an adjustment is made.

The last adjustment is in the keyer. Adjust the potentiometer R_1 to the point where the oscillator cannot be heard between dots and dashes at normal keying speed.

An All-Purpose 813 Amplifier

Figs. 6-59 through 6-62 show the circuit and photographs of an 813 amplifier designed for c.w., a.m., or s.s.b. operation. Provision has been made for convenient changing from one mode to another as well as to any of the bands from 80 through 10 meters.

The circuit is shown in 6-60. A turret-type grid circuit is used and the output circuit is a pi network designed to work into coax cable. The inductor is the rotary-type variable. Provision for neutralizing is included. R_1 is a parasitic suppressor.

For Class C c.w. or phone operation, S_4 is open. The 90 volts of fixed bias, furnished by a small bias supply and regulated by the VR90, is augmented by a drop of about 50 volts across the grid-leak resistor R_2 at a normal grid current of 15 ma. This brings the total bias to 140 volts. With S_4 closed, the grid leak is short-circuited and the 90 volts of fixed bias alone remains for AB_2 s.s.b. operation. (An advantage in AB_2 for c.w. operation is that it preserves the keying characteristics of the exciter better than with Class C operation.) R_3 should be adjusted so that the VR90 just ignites with no excitation.

Screen voltage is regulated at 750 volts by a string of five 0A2s for s.s.b. operation. When the grid drive is increased for Class C operation, the screen current increases, increasing the drop across the screen resistor R_5, and the screen voltage falls to 400. The regulators then lose control and the amplifier is ready for plate-screen modulation.

The screen is protected against excessive input, should the load or plate voltage be removed, by the overload relay K_1. The tripping point is set at 40 ma. by the variable shunt resistor R_4. If the relay trips, current through R_6 will hold the screen circuit open until plate voltage is removed. One meter, M_1, measures cathode current, while the other meter, M_2, may be switched to read either grid current or screen current.

Forced-air ventilation is always advisable for a medium- or high-power amplifier if it is buttoned up tight to suppress TVI. A surplus 100 c.f.m. blower does the job more than adequately.

Construction

The amplifier is built on a 13 × 17 × 4-inch aluminum chassis fastened to a standard 12¾ × 19-inch rack panel. The r.f. output portion is enclosed in a 12½ × 13 × 8½-inch box made of aluminum angle and sheet. The VR tubes, relay, blower and meters are mounted external to the box.

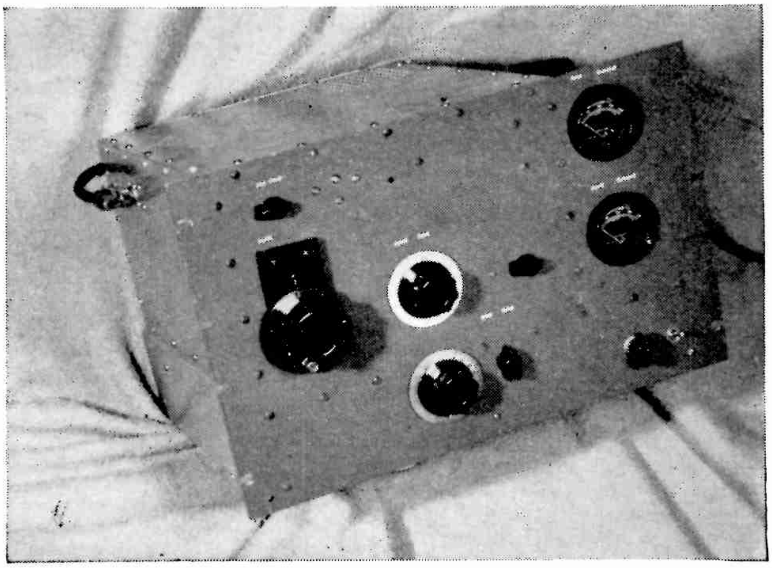
The grid tank-circuit components are mounted underneath the chassis and are shielded with a 5 × 7 × 3-inch aluminum box. A standard chassis of these dimensions might be substituted. The bias and filament transformers are in a second box measuring 6 by 3 by 3 inches. This type of construction, together with the use of shielded wire for all power circuits, was followed to reduce TVI to a minimum. Each wire was bypassed at both ends with 0.001-μf. ceramic disk capacitors. L_4 can be adjusted to series resonate with the 600-$\mu\mu$f. capacitor at the frequency of the most troublesome channel. A Bud low-pass filter completes the TVI treatment. As a result, the amplifier is completely free of TVI on all channels even in most fringe areas.

Adjustment

In the pi network, the output capacitors are fixed. However, the adjustment of the network is similar to that of the more conventional arrangement using a variable portion of the output capacitance. The only difference is that the "fine" loading adjustment is done with the variable inductor.

The inductor is fitted with a Groth turns counter, making it easy to return to the proper

Fig. 6-59—W4SUD's all-purpose 813 amplifier. The output-capacitor switch (coarse loading) is above the turns counter for the variable inductor. Dials near the center are for the plate tank capacitor C_4 (above) and the grid tank capacitor C_1 (below). To the right of the dials are the controls for the plate padder switch S_3 (above) and the grid band switch S_1 (below). The toggle switch below the meters is the mode switch S_4 with the meter switch S_5 to the left. Ventilating holes are drilled in the cover in the area above the tube. The output connector is on the left-hand wall of the shielding box



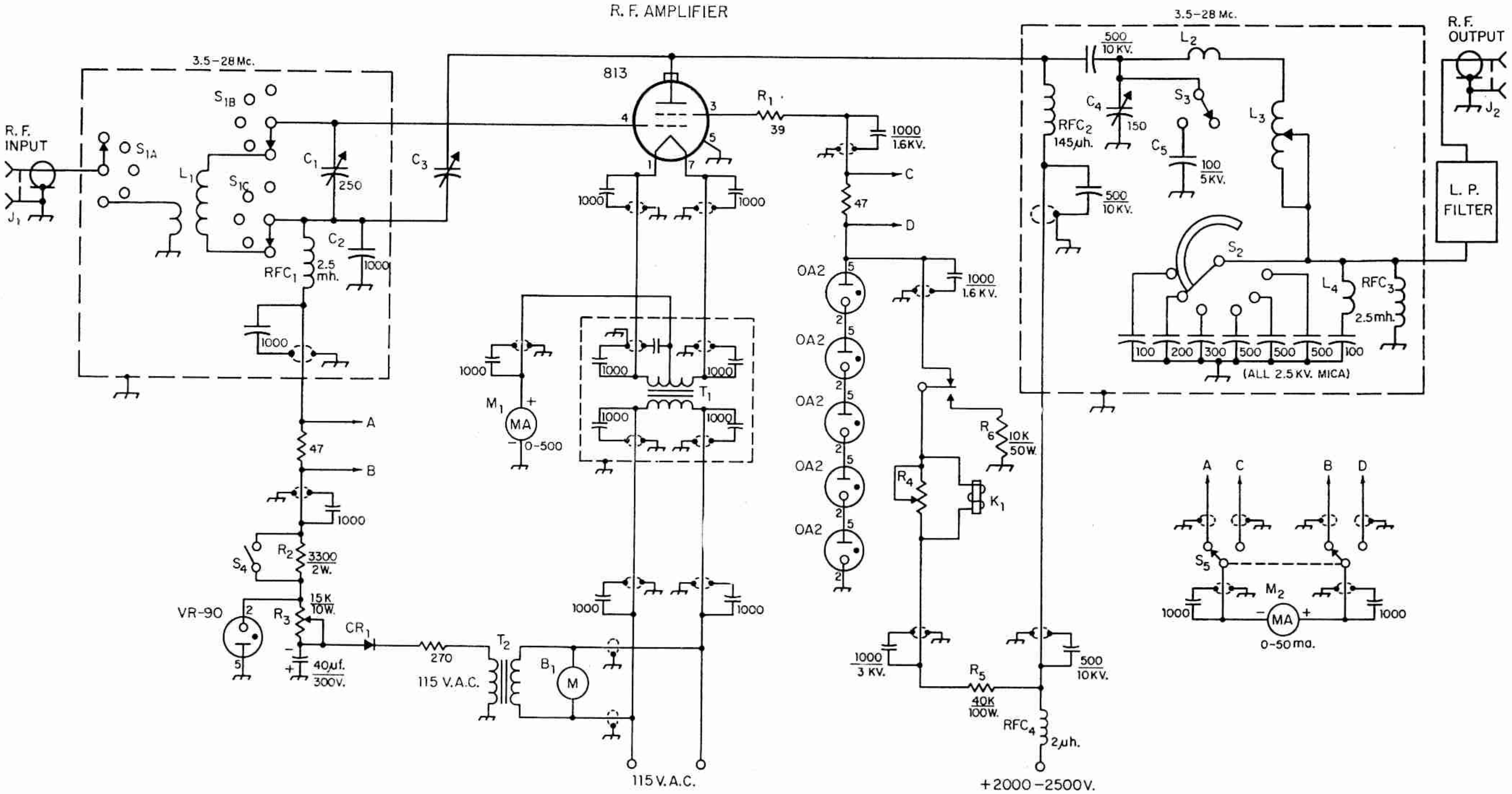


Fig. 6-60—Circuit of the all-purpose 813 amplifier. Unless otherwise designated, capacitances are in $\mu\mu$f. Capacitors marked with polarity are electrolytic. Other capacitors not listed below should be ceramic. Resistances are in ohms.

B_1—Ventilating blower, 100 c.f.m. (surplus).
C_1—250-$\mu\mu$f. variable (Hammarlund MC-250-M).
C_2—1000-$\mu\mu$f. mica.
C_3—Neutralizing capacitor, 10 $\mu\mu$f. maximum (Johnson 159-250).
C_4—150-$\mu\mu$f. 6000-volt variable (Johnson 153-12).
C_5—100-$\mu\mu$f. 5000-volt fixed capacitor (surplus vacuum Amperex VC-100, or two 200-$\mu\mu$f. 5000-volt micas in series).
CR_1—130-volt 50-ma. selenium rectifier.
J_1, J_2—Coaxial receptacle (SO-239).
K_1—Screen overload relay, 2500 ohms, 7 ma. (Potter & Brumfield KCP5).
L_1—3.5 Mc.—32 turns No. 20, 1-inch diam., 2 inches long, 5-turn link (B&W 3015 or Airdux 816).
—7 Mc.—18 turns No. 20, ¾-inch diam., 1⅛ inches long, 3-turn link (B&W 3011 or Airdux 616).
—14 Mc.—10 turns No. 18, ⅝-inch diam., 1¼ inches long, 2-turn link (B&W 3006 or Airdux 508).
—21 Mc.—7 turns No. 18, ⅝-inch diam., ⅞ inch long, 1-turn link (B&W 3006 or Airdux 508).
—28 Mc.—5 turns No. 18, ⅝-inch diam., ⅝ inch long, 1-turn link (B&W 3006 or Airdux 508).
L_2—3 turns 3/16-inch copper tubing, 1-inch diam., 1¾ inches long.
L_3—15-μh. variable inductor (B&W 3852).
L_4—See text.
M_1, M_2—3½-inch d.c. milliammeter.
R_1—39 ohms, ½-watt carbon.
R_2—3300 ohms, 2 watts.
R_3—15,000 ohms, 10 watts with slider.
R_4—2000-ohm 4-watt variable resistor (Mallory M2-MPK).
RFC_1, RFC_3—2.5-mh. r.f. choke (National R-50 or similar).
RFC_2—Plate r.f. choke (National R-175-A).
RFC_4—V.h.f. choke (National R-60).
S_1—Rotary switch: 3 wafers, 3 poles, 11 positions per pole, 5 positions used (Centralab PA-0 wafers, PA-301 index).
S_2—Rotary switch: single pole, 10 positions, progressively shorting, 6 positions used (Centralab PA-2042).
S_3—Rotary switch: s.p.s.t., ceramic (antenna link switch from BC-375 tuning unit, or Communications Products Model 65).
S_4—S.p.s.t. toggle switch.
S_5—D.p.d.t. rotary switch (Centralab 1405).
T_1—Filament transformer: 10 volts, 5 amp. (Thordarson 21F18).
T_2—Bias transformer: 120 volts, 50 ma.; 6.3 volts, 2 amp., filament winding not used; could be used for pilot light (Merit P-3045).

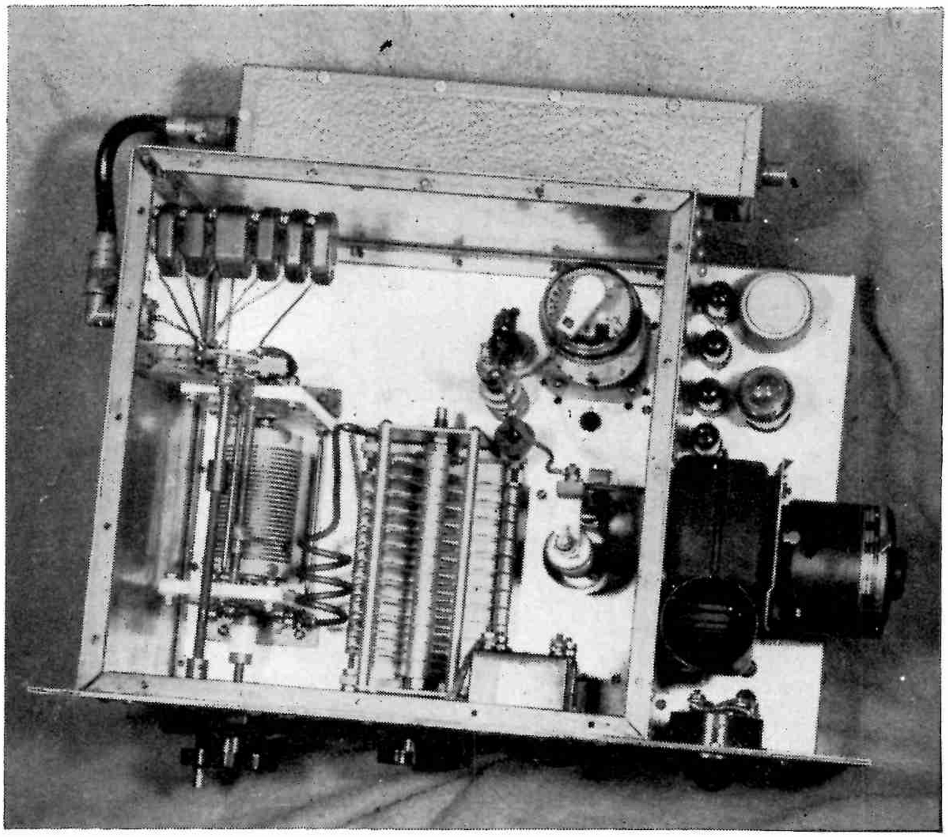


Fig. 6-61—This view shows the placement of components on the chassis. The 813 socket is mounted on spacers over a large clearance hole in the chassis. The several mica output capacitors are assembled in a stack on a threaded rod fastened to the left-hand wall of the shielding box. The neutralizing capacitor and the 80-meter plate padder are to the right of the tank capacitor. To the right of the box are the five 0A2s (the front one hidden), the screen overload relay and the VR90, the blower and meters.

setting for each band. Until the settings for each band have been found, S_3 should be turned so that all of the output capacitance is in circuit. The inductor should be set near maximum for 80, and approximately half maximum for 40. On the higher-frequency bands, the inductor should be set so that the circuit resonates with the tank capacitor near minimum capacitance. Loading should increase as the output capacitance is decreased. A change in output capacitance requires a readjustment of C_4 for resonance. When the loading is near the desired point, final adjustment can be made by altering the inductance slightly.

A 20-A or similar exciter is well suited as a driver for this amplifier on all modes. The 813 runs cool at 500 watts input on a.m. and c.w. and at 1000 watts p.e.p. on s.s.b. (Originally described in *QST* for August, 1958.)

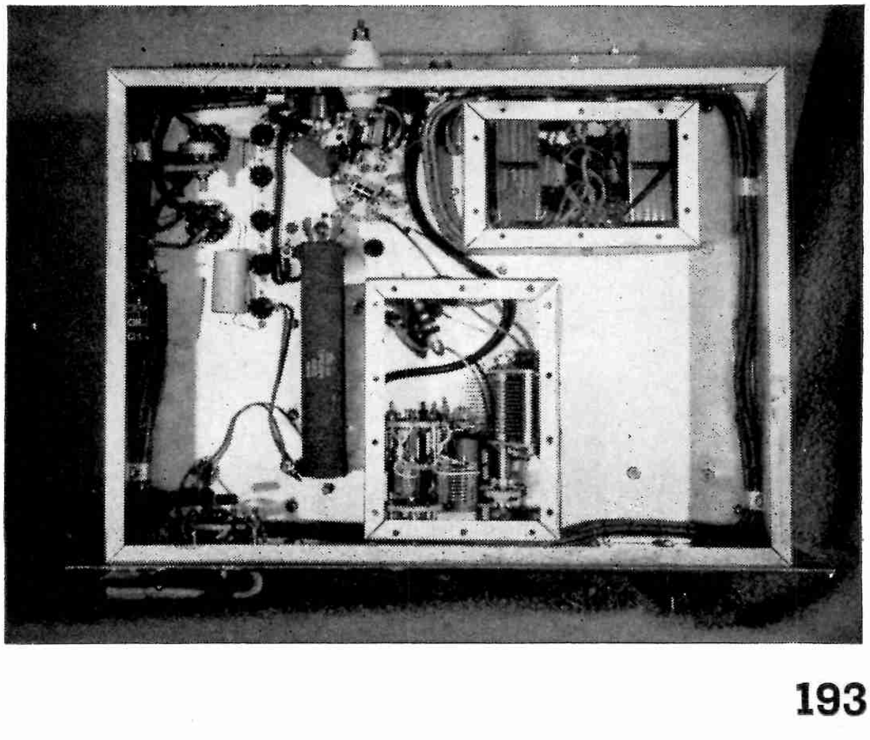


Fig. 6-62—Bottom view of the all-purpose 813 amplifier. The grid tank-circuit components within dashed lines in Fig. 6-70 are enclosed in the box at lower center. Input links are wound over ground ends of grid coils. Filament and bias transformers are in the second box. The large resistor to the left of the grid box is the screen resistor. The variable resistor in the upper left-hand corner is the relay shunt R_4. The selenium bias rectifier is fastened against the left-hand wall of the chassis.

A Medium-Power Tetrode Amplifier

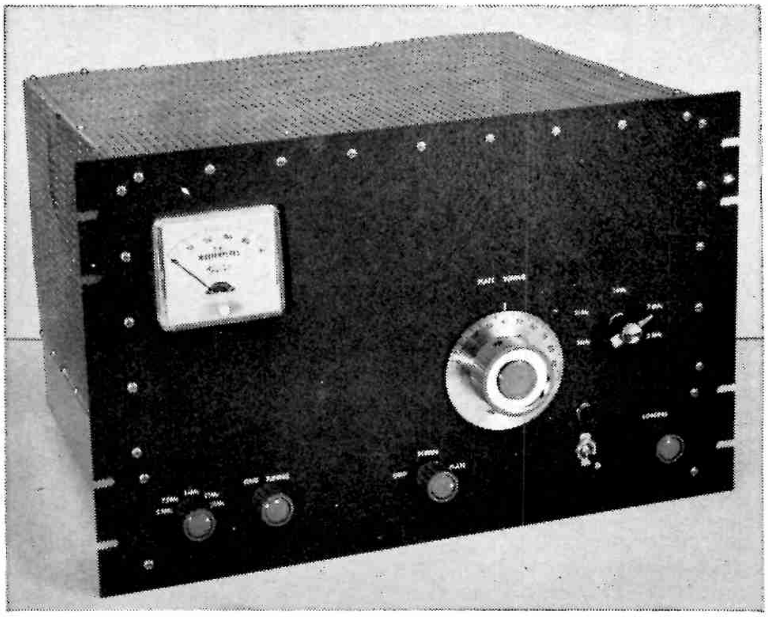


Fig. 6-63—This medium-power tetrode amplifier is assembled on a 17 × 12 × 3-inch aluminum chassis with a 19 × 12¼-inch rack panel. Controls along the bottom of the panel are for the grid band switch, grid tuning capacitor, meter switch, a.c. power, and pi-network loading capacitor. Above are the controls for the plate tank capacitor and plate band switch. The sides and back of the shielding enclosure are a single piece of Reynolds perforated aluminum sheet "wrapped" around the chassis. A 1-inch lip is bent along the three top edges so that the top cover can be fastened on with sheet-metal screws.

Figs. 6-63 through 6-66 show photographs and circuit diagram of an amplifier using an RCA 7094 tetrode that will handle up to 500 watts input on c.w. or 330 watts with plate-screen modulation. Construction has been simplified by the use of manufactured subassemblies — a Harrington Electronics GP-50 multiband grid tank and a B & W type 851 bandswitching pi-network inductor. The amplifier is neutralized by the capacitive-bridge method. R_1 and L_5 are adjusted to suppress v.h.f. parasitic oscillation. The single milliammeter M_1 may be switched to read either grid or plate current. The shunt R_2 multiplies the original 50-ma. scale by 10, giving readings up to 500 ma. when the meter switch S_3 is in the plate-current position. Forced-air ventilation is provided by a small blower B_1. Shielded wire is used in all power circuits and terminal leads are bypassed for v.h.f. as they enter the chassis.

Construction

The plate blocking capacitor is threaded onto one of the plate tank-capacitor stator rods. Plate-circuit leads are made of ½-inch copper strip. Screen and filament bypasses are connected directly between the tube-socket terminals and the perforated sheet. Each of the three screen terminals is bypassed with a 1000-$\mu\mu$f. 1600-volt disk ceramic capacitor. The grid-tank unit is spaced from the front wall of the chassis on 1-inch pillar insulators to provide space for an insulating shaft coupling.

Along the rear wall of the chassis are the coax

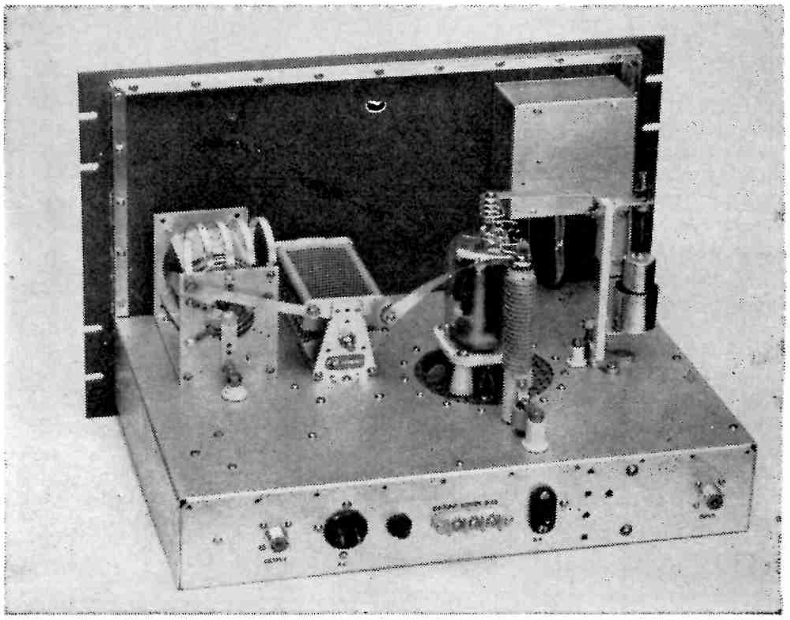


Fig. 6-64—Rear view of the medium-power amplifier. The shafts of the plate band switch and plate tuning capacitor are 2¾ and 6¼ inches from the left-hand end of the chassis in this view. A ventilating hole somewhat larger than the tube socket (829-B type) is centered 6½ inches from the right-hand end of the chassis and 6 inches from the rear. A piece of perforated aluminum covers the hole and supports the tube socket mounted on 1-inch ceramic cones. Feed-through insulators carry connections to the bottom terminals of the plate tank-coil unit, the plate r.f. choke and the neutralizing capacitor. The meter is enclosed in a 4 × 4 × 2-inch aluminum box.

Medium-Power Tetrode Amplifier

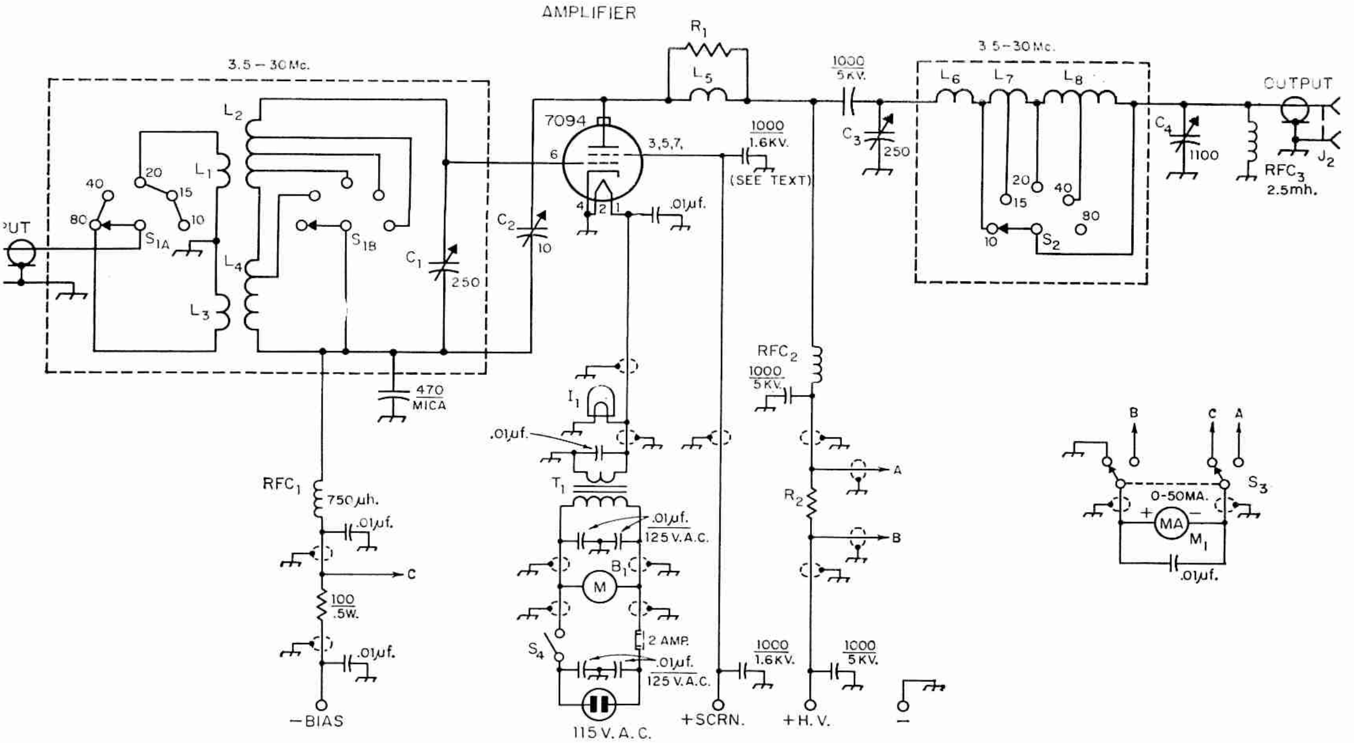


Fig. 6-65—Circuit of the 7094 amplifier. Unless specified otherwise, capacitances are in μμf. All fixed capacitors rated at less than 5 kv. are disk ceramic. The 5-kv. capacitors are TV-type ceramics (Centralab 858). Dashed lines in grid circuit enclose components of Harrington GP-50 multiband tank unit. Those in the plate circuit enclose components of the B & W 851 pi-network inductor.

B_1—Blower (Allied Radio Cat. No. 72P715).
C_1—250-μμf. midget variable (special).
C_2—Neutralizing capacitor—11 μμf. max. (Johnson N125).
C_3—250-μμf. 3000-volt variable (Johnson 250E30).
C_4—1100-μμf. variable—triple-gang broadcast replacement type, 365 μμf. (or more) per section, sections connected in parallel.
I_1—6.3-volt dial lamp.
J_1, J_2—Coax receptacle (SO-239).
L_1—2 turns No. 16, 1 inch diam., over ground end of L_2.
L_2—14 turns No. 16, ¾ inch diam., 2 inches long.
L_3—3 turns No. 16, 1 inch diam., over ground end of L_4
L_4—38 turns No. 22, ¾ inch diam., 1½ inches long.
L_5—3 turns No. 12, ⅜ inch diam., 1 inch long.
L_6—4 turns 3⁄16 × 3⁄16-inch copper strip, 1⅜ inches diameter, 2½ inches long.
L_7—4¾ turns No. 8, 2½ inches diam., 1¾ inches long, tapped at 3 turns from the L_8 end.
L_8—9½ turns No. 12, 2½ inches diam., 1½ inches long, tapped at 6 turns from the output end (see text).
Note: L_7 and L_8 are mounted close together on the same axis; L_6 is mounted at right angles.
M_1—D.c. milliammeter, 0–50-ma. scale—3⅜-inch rectangular (Triplett Model 327-PL).
R_1—Three 150-ohm 1-watt carbon resistors in parallel.
R_2—Approx. 32 turns No. 24 on a ¼-inch diam. form (see measurements section for method of adjustment).
RFC_1—750-μh. r.f. choke (National R-33).
RFC_2—Plate r.f. choke 120μh (Raypar RL-101).
RFC_3—2.5-mh. r.f. choke (National R-50).
S_1—Two-wafer 5-position ceramic rotary switch.
S_2—Special heavy-duty 5-position rotary switch (component of B & W inductor unit).
T_1—Filament transformer: 6.3 volts, 3.5 amps. minimum (Thordarson 21F11).

output connector, a.c. power connector, fuse, screen-voltage, bias and ground terminals, high-voltage connector (Millen) and the coax input connector. Strips of ½-inch aluminum angle fastened to the panel provide a means of fastening the shielding enclosure to the panel. Paint should be removed where the angle rests against the panel so that there will be good electrical contact between the two.

Preliminary Adjustment

To maintain a tank Q of 10 at 4 and 7.3 Mc., 4 turns should be removed or shorted out at the front end of the B&W unit, and the 40-meter tap should be moved one turn toward the rear. (For operation at less than maximum ICAS ratings, see pi-network charts earlier in this chapter.)

Before applying excitation, the amplifier should be checked for v.h.f. parasitic oscillation as described earlier in this section. A resistor of about 20,000 ohms should be connected between the bias terminal and ground. Full plate voltage may be applied, but the screen should be operated from an adjustable 50,000-ohm 50-watt series resistor connected to the plate supply. The grid band switch should be turned to the 10-meter position and the plate switch to the 80-meter position. With the meter switched to read plate current, the screen resistance should be reduced until the plate power input is about 100 watts. The meter should then be switched to read grid current and the recommended procedure followed. The objective is to suppress the parasitic oscillation with the smallest possible coil to keep the parasitic-circuit

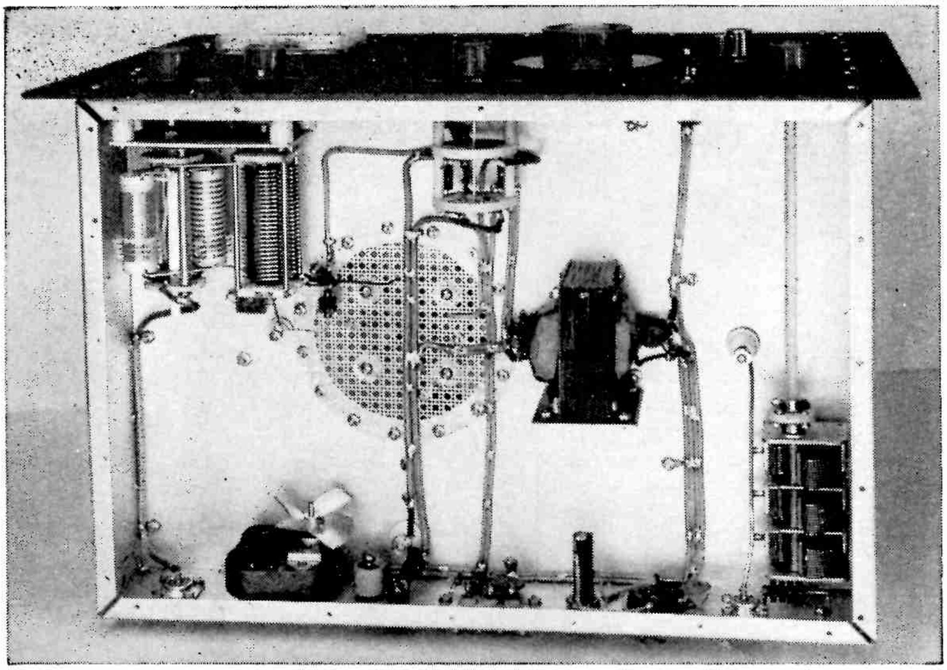


Fig. 6-66—Bottom view of the 7094 amplifier. The grid-tank assembly in the upper left-hand corner and the output loading capacitor in the lower right-hand corner are placed so that the shaft of the latter and the shaft of the grid band switch are 1½ inches from the ends of the chassis. Spacers between the chassis and the output capacitor bring its shaft level with those of the grid-tank unit. The meter switch is at the center. The filament transformer is mounted on an aluminum bracket. The ventilating fan is bolted against the rear wall of the chassis.

resonant frequency between the two v.h.f. TV bands. If oscillation is detected, additional loading resistors should be tried first. If this does not work, another turn should be added to the coil, or the turns squeezed closer together. With the parasitic coil described, the resonant frequency of the circuit is about 100 megacycles.

Neutralizing

Neutralizing should be done with excitation applied to produce rated grid current. The input and output circuits should be tuned to the same frequency. Plate and screen voltages should be disconnected at the transmitter terminals. The neutralizing capacitor should then be adjusted until a point is found where there is no change in grid current as the plate tank circuit is tuned through resonance. The output capacitor should be set at maximum capacitance for this check. After plate and screen voltages have been applied and the amplifier loaded, the neutralizing capacitor should be given a final adjustment to the point where minimum plate current and maximum grid and screen currents occur simultaneously.

Power Supply

Maximum ICAS ratings on the 7094 are 1500 volts, 330 ma. on c.w., 1500 volts, 200 ma. (max.) Class AB_1 s.s.b., and 1200 volts, 275 ma. for a.m. phone. However, the tube will work well at plate voltages down to at least 700 volts, provided appropriate values are used in the pi network as mentioned previously. The recommended screen voltage is 400 for all classes of operation at screen currents up to 30 ma., depending on the type of operation. Therefore a regulated screen voltage can be obtained using a pair of 0D3s and one 0C3 in series. If screen voltage is obtained from the plate supply, an adjustable 100-watt 75,000-ohm series resistor should be used and the value adjusted to obtain the desired operating plate current after initial tuning adjustments have been made.

Biasing

A fixed biasing voltage of 50 is required for s.s.b. operation. Batteries should last indefinitely. The biasing voltage may also be obtained from a voltage divider across a VR tube with suitable series resistor. A biasing voltage of 130 is recommended for plate-modulated Class C service, and 100 volts for c.w. operation. Recommended grid current is 5 ma. If the screen is operated from a fixed-voltage source, a source regulated by an 0A3 should provide plate-current cut off. The balance of the required operating bias may be obtained from a grid leak (5000 ohms for c.w. or 11,000 ohms for phone). In case the screen is supplied through a dropping resistor from the plate supply, fixed biasing voltages of 100 for c.w. or 130 for phone (no grid leak) should provide reasonable protection for the tube in case of failure of excitation.

The rated driving power is 5 watts, easily furnished by a 2E26 without pushing it. Existing transmitters using a 6L6, 6146 or 807 in the final may be used if provision is made for controlling the output of these units by adjustment of screen voltage.

A Grounded-Grid Half Kilowatt

The amplifier shown in Figs. 6-67, 6-69 and 6-70 will run at about 500 watts input on c.w. — or p.e.p. input as an s.s.b. linear — on all bands from 80 through 10 meters. The unit is small enough to sit on the operating table right along with the rest of the station equipment; no need for big racks here.

Using a pair of 811As in parallel in the grounded-grid circuit, this rig is a good one to use following transmitters such as the Viking Ranger, DX-40, Globe Scout, and others of similar power class, for a worth-while increase in power output on c.w. As a linear amplifier following an s.s.b. exciter it requires no swamping because the 811A grids provide a fairly constant load in themselves, and also the fed-through power with grounded-grid presents an additional constant load to the driver. The total driving power needed on any band is less than 20 watts.

An additional useful feature is a built-in directional coupler using a version of the "Mickey Match." Besides its obvious application for checking the s.w.r. on the transmission line to the antenna or for help in tuning up a coax-coupled antenna coupler, it is practically indispensable as an indicator of relative power output in tuning the amplifier.

The Circuit

A number of tube types could be used in an amplifier of this power class, but the 811As are a good choice because they do not need a bias supply and are not expensive. (Surplus 811s can be used if you don't want to buy new tubes; the ratings are not quite as high but they can be pushed a bit in intermittent service such as c.w. and s.s.b.)

The complete circuit is shown in Fig. 6-68. To save trouble and work, standard components are used throughout — the only special construction is the shielding and a few simple r.f. chokes. The tube filaments are driven directly from coax input from the driver; no tuning is used or is needed in this circuit. The filaments are kept above ground by the B & W type FC15 filament choke.

The plate tank is the familiar pi network, using a B & W type 851 tapped coil and band-switch assembly. This assembly has been modified slightly in two respects: First, the copper-strip 10-meter coil normally mounted at the top of the rear plate is taken off and moved so that it is supported between the tank assembly and the stator of the tank tuning capacitor as shown in Fig. 6-69. A short length of copper strip is bolted between the free end of the coil and the right-hand stator connection of the tuning capacitor, to support the free end. This change is made in order to avoid the long lead that would have to be run from the capacitor to the regular input terminal on the tank assembly, since this terminal is at the right-hand side of the assembly as viewed from the top. The turns of the 10-meter coil are also squeezed together a bit to increase the inductance, because it was found that a rather large amount of capacitance had to be used to tune the circuit to the band with the coil at its original length. The length is now 1⅝ inches between mounting holes.

The second modification is the addition of a pair of switch contacts on the rear switch plate of the tank assembly. There is an extra position on this plate with holes already provided for contacts, and the additional set of contacts is used to switch in fixed output loading capacitance on 80 meters, where a large output capacitance is needed. The variable loading capacitor, C_3, with the five fixed mica capacitors, C_5 to C_9 inclusive, give continuous variation of capacitance up to 1275 $\mu\mu$f. on all bands, including the regular switch position for the 80-meter band. However, if the switch is turned to the extra position an additional 1000-$\mu\mu$f. mica capacitor is connected in parallel, so that continuous variation of capacitance to over 2200 $\mu\mu$f. is possible on 80. This takes care of cases where the load resistance

Fig. 6-67—This amplifier operates at a plate input of approximately 500 watts, uses a pair of 811As in grounded-grid, and is complete with power supply on a 13 × 17 × 4-inch chassis. The rack panel is 10½ by 19 inches. Front-panel controls include the plate tuning capacitor and band switch in the center, filament and plate power switches with their pilot lights at the lower left, sensitivity control and forward-reflected power switch for the directional coupler at the lower right, variable loading capacitor and auxiliary loading-capacitor switch underneath the 0-1 milliammeter at the right, and the grid-cathode milliammeter with its switch at the upper left. The filter choke, 866As and plate transformer occupy the rear section of the chassis.

happens to be unusually low or reactive.[1]

A 500-ma. d.c. meter is used for reading either total cathode current or grid current alone. The cathode current is read in preference to plate current because of safety considerations. Putting the meter in the hot d.c. plate lead leaves nothing but a little plastic insulation between the high voltage and the meter adjusting screw. It is a bit of a nuisance to have to subtract the grid current from the cathode current in order to find the plate current, but it isn't serious. The d.c. grid circuit has a jack, J_3, for introducing external bias either for blocked-grid keying or for cutting

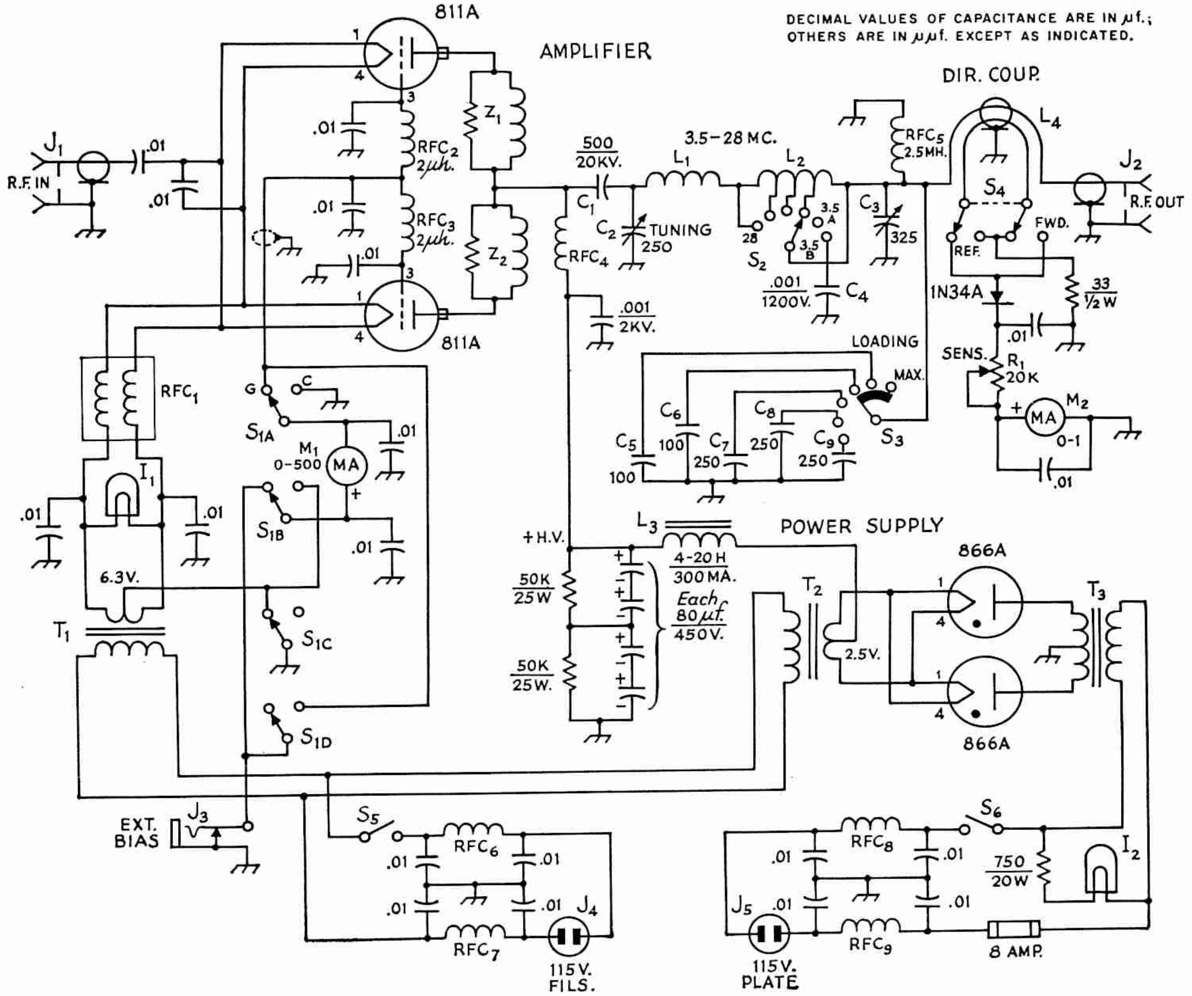


Fig. 6-68 — Circuit diagram of the parallel-811A grounded-grid amplifier. Unless otherwise specified, fixed capacitors are disk ceramic, 600-volt rating.

C_1—500 μμf., 20,000 volts (TV "doorknob" type).

C_2—250-μμf. variable, 2000 volts (Johnson 250E20).

C_3—325-μμf. variable, receiving type (Hammarlund) MC-325-M).

C_4–C_9, inc.—1200-volt mica, case style CM-45.

I_1, I_2—6.3-volt dial lamp, 150-ma. (No. 47).

J_1, J_2—Coax connector, chassis mounting.

J_3—Closed-circuit phone jack.

J_4, J_5—115-volt male connector, chassis mounting (Amphenol 61-M1).

L_1, L_2, S_2—5-band pi-network coil-switch assembly; see text (B & W 851).

L_3—Swinging choke, 4-20 henrys, 300 ma. (UTC S-34).

L_4—Section of coax line with extra conductor inserted; see Footnote 1 for construction references.

M_1, M_2—Milliammeter, 3½-inch plastic case (Triplett 327-PL).

R_1—20,000-ohm composition control, linear taper.

RFC_1—*Filament-choke assembly, to carry 8 amp.* (B & W FC15).

RFC_2, RFC_3—2 μh. (National R-60).

RFC_4—90 μh.; 4⅜-inch winding of No. 26, 40 t.p.i., on ¾-inch ceramic form (B & W 800).

RFC_5—2.5 mh., any type.

RFC_6–RFC_9, Incl.—18 turns No. 14 enam., close-wound, ½-inch diam., self-supporting.

S_1—4-pole 2-position rotary, nonshorting (Mallory 3242J or Centralab 1450).

S_2—Part of tank assembly; see L_1 L_2.

S_3—Miniature ceramic rotary, 1 section, 1 pole, 6 positions used, progressive shorting (Centralab 2042).

S_4—Miniature ceramic rotary, 1 section, 2 poles, 2 positions used, nonshorting (Centralab 2003).

S_5, S_6—S.p.s.t. toggle.

T_1—Filament transformer, 6.3 volts, 8 amp. min. (UTC S-61).

T_2—Filament transformer, 2.5 volts, 10 amp. (UTC S-57).

T_3—Plate transformer, 3000 volts center-tapped, 300 ma. d.c. (UTC S-47).

---

[1] These contacts can be obtained directly from the manufacturer of the tank assembly. To secure a set of contacts with mounting hardware, send one dollar to Barker & Williamson, Beaver Dam and Canal, Bristol, Penna., specifying the type of tank assembly for which they are wanted. The contacts are not catalog items and are not available through dealers.

A 90-Watt Amplifier

Construction

Most of the components can be identified in Figs. 6-47, 6-48 and 6-50, but a few construction notes are in order. The octal socket for the 6146 is mounted on two ½-inch-long collars above the usual 1⅛-inch diameter hole in the chassis. The three .001-μf. ceramic capacitors connected to the cathode pins (1, 4 and 6) ground to the chassis at lugs under the nuts holding the socket-mounting screws. The .001-μf. ceramic capacitors in the screen and heater circuits ground to their respective wire shields which in turn are connected to the same ground lugs as the cathode circuit. The grounded side of the 680-μμf. capacitor in the grid-circuit return should also be soldered to one of the ground lugs.

The neutralizing capacitor, C_2, has its rotor insulated from the chassis by mounting it in extruded fiber washers and a suitable hole in the chassis. Connection to the rotor should be made under the chassis by using a suitable soldering lug under the nut on the threaded sleeve bearing. (Old volume controls are a good source for this lug.)

The high-voltage lead from the base of RFC_3 is run in well-insulated wire to a feed-through bushing that runs through the chassis and to the meter switch terminal z_1. A high-voltage bypass capacitor is connected between the bushing and the chassis.

A simple clamp, Fig. 6–52, holds the length of RG-58/U from C_4 in place and at the same time insures that the r.f. leaves the compartment via the inside of the cable and not the outside.

Aluminum cane metal is available in many hardware stores, and it is an easy matter to bend

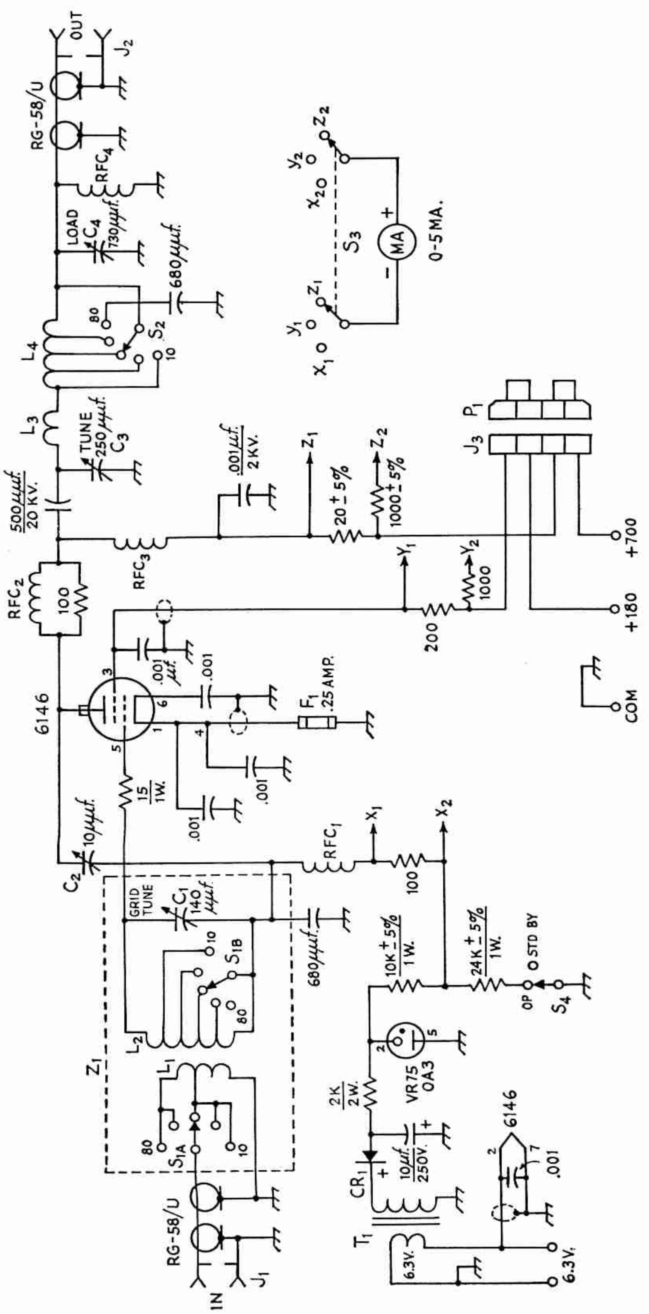


Fig. 6-49—Circuit diagram of the all-purpose amplifier and its bias supply. Unless otherwise indicated, resistors are ½ watt.

C_1—140-μμf. midget variable (Hammarlund APC-140-B).
C_2—10-μμf. midget variable (Hammarlund HF-15X with one stator plate removed).
C_3—250-μμf. variable (Hammarlund MC-250M).
C_4—730-μμf. variable (Broadcast receiver replacement, 365 μμf. each section, connected in parallel).
CR_1—20-ma. 130-volt selenium rectifier.
J_1, J_2—Coaxial cable connector, SO-239.
J_3—4-pin tube socket.
L_1—3¾ turns No. 18 at grid end of L_2, tapped 2 turns from ground end.
L_2—50 turns No. 24, 1¾ inches long on ¾-inch diameter threaded ceramic form. Tapped at 5, 8, 13 and 25 turns from grid end.
L_3—4¼ turns No. 14, 1³⁄₁₆ diam., ⅝ inch long.
L_4—18 turns No. 16, 2-inch diameter, 10 t.p.i. Tapped at 1⅛, 5⅛ and 11⅛ turns from plate end. (B&W 3907-1).
P_1—4-prong plug, with jumper connections as shown.
RFC_1—2.5-mh. 100-ma. r.f. choke (National R-50).
RFC_2—5 turns No. 16 wire, wound on 100-ohm 1-watt resistor.
RFC_3—1-mh. 500-ma. r.f. choke (Johnson 102-752).
RFC_4—2.5-mh. 125-ma. r.f. choke (National R-100S).
S_1—2-pole 6-position (5 used) miniature ceramic switch (Centralab PA-2002).
S_2—1-pole 6-position (5 used) ceramic switch (Centralab 2501).
S_3—2-pole 6-position (5 used) non-shorting miniature ceramic switch. (Centralab PA-2003). Alternate contacts used only, to increase voltage rating.
S_4—S.p.s.t. toggle switch.
T_1—6.3-volt filament transformer (Stancor P-6134).
Z_1, comprising C_1, L_1, L_2 and S_1 is Harrington Electronics GP-20 unit. Capacitors showing polarity are electrolytic; 680-μμf. capacitors are silver mica, .001-μf. are ceramic.

a piece of it to form the cover. Make the cover with lips on the vertical portion that slip tightly over the sides of the Minibox, and with a bend at the bottom that can be fastened to the chassis. Another piece of cane metal should be cut to serve as a bottom cover; mounting the chassis on rubber feet lifts it above the table and permits good air circulation through the unit.

The self-supported inductor L_3 can be wound on the envelope of one of the 6DE4 rectifiers, removed and pulled apart slightly to give the specified winding length. The taps on L_4 are made by first bending inward the wire on either side of the turn to be tapped, then looping the tap wire around the turn and soldering it securely in place. Both L_3 and L_4 are supported only by their leads.

Testing and Adjustment

With all tubes in their sockets except the 6146, the line cord should be plugged in and the power switch turned on. The bias-supply 0A3 should glow immediately and the rectifier filament and heaters should light up. The screen-supply regulators should glow. If a voltmeter is available, the high-voltage supply should show first around 400 volts, and then rise slowly to about 950 volts. Switch off the power; the plate supply voltage should decay to less than 100 in under 20 seconds, indicating that the 40,000-ohm resistors are "bleeding" the supply. Note also how long it takes for the voltage to reach a value of only a few volts: this will demonstrate forcefully how long it takes to discharge a high-capacitance filter.

When the power supply has discharged, plug in the 6146, connect the plate cap, and set S_4 to STAND BY. Set the neutralizing capacitor C_2 at half capacitance and the band switches on 80 meters. Turn on the power and set the meter switch, S_3, to read plate current. The 6146 heater should warm up. Now flip S_4 to operate; the meter should read 10–20 ma. (.2-.4 on the scale). Switching to read screen current, the meter should show under 1 ma. (2 divisions on the meter). There should be no grid current.

Turn off the power and remove the three rectifier tubes. Connect at J_1 the driver or excita-

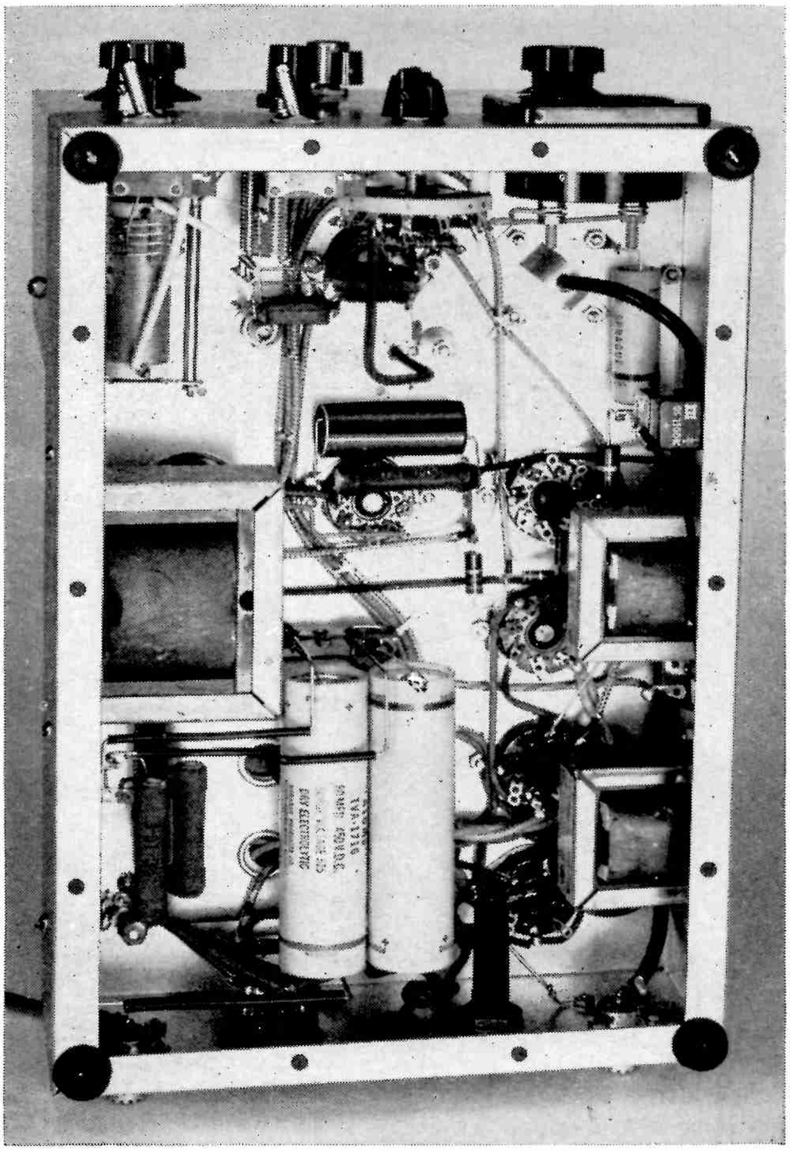


Fig. 6-50—Bottom view of the all-purpose amplifier. The 150-ma. filter choke is mounted on the left-hand wall; the smaller filter choke, the small filament transformer (T_1 in Fig. 6-51) and the selenium rectifier are mounted on the right-hand wall. The strap of aluminum, visible below the meter at the top right, provides additional support for the length of RF-58/U cable that runs to the output coaxial connector. All power leads except the high voltage to the plate are run in shielded wire.

off the plate current during receiving, and a four-pole switch, S_1, is therefore needed for handling the meter switching while keeping all circuits functioning normally.

The power supply uses 866As with a plate transformer giving 1500 volts each side of the center tap, and working into a single-section choke-input filter. The filter capacitor consists of four 80-μf. electrolytics connected in series to handle the voltage, giving an effective filter capacitance of 20 μf. This supply is running well below its capabilities in the intermittent type of operation represented by c.w. and s.s.b., and the amplifier is somewhat "over-powered" in this respect. A lighter plate transformer can be used since the average current in regular operation is only about half the maximum tube rating of 350 ma. for the pair.

The a.c. inputs to both filaments and plates have TVI filters installed right at the a.c. connectors. The chokes in these filters, RFC_6 to RFC_9 inclusive, are homemade by winding 18 turns of No. 14 enameled wire close-wound on a half-inch dowel or drill.

Construction

The only space available for the filament transformers is below chassis, so these are mounted on the front wall of the chassis as shown in Fig. 6-70. There is plenty of room for all other power-supply parts below chassis, and the photographs make any further comment on this section unnecessary.

The r.f. layout shown in Fig. 6-69 is almost an exact copy of the circuit layout as given in Fig. 6-68. The plate blocking capacitor, C_1, is mounted on a small right-angle bracket fastened to the left-hand stator connection of the tank capacitor, C_2. The tube plates are connected to C_1 through individual parasitic-suppressor assemblies, Z_1 and Z_2. The hot end of the plate choke, RFC_4, also connects to this same point. The tank capacitor is mounted on ¾-inch ceramic pillars to bring its shaft to the same height as the switch shaft on the tank-coil assembly. The capacitor is grounded by connecting the bottom of its frame through a half-inch wide strip of aluminum to essentially the same point at which the plate-choke bypass capacitor, a 0.001-μf. 2000-volt disk, is grounded. The ground end of the aluminum strip actually is under the bottom of the plate choke, and the ground lug for the bypass capacitor is just to the left. This strip, plus short leads in the circuit from the tube plates through the tank capacitor to ground, keep the resonant frequency of the loop thus formed well up in the v.h.f. region; this is important because it permits using low-inductance parasitic chokes in shunt with the suppressor resistors, and thus tends to keep the r.f. plate current at the regular operating frequencies out of the resistors. With other tank grounding arrangements originally tried, larger parasitic chokes had to be used and it was impossible to prevent the resistors from burning up when operating on 10, 15 and even 20 meters. Now they do not overheat on any frequency, and v.h.f. parasitics are nonexistent — although without the suppressors the parasitics are only too much in evidence.

The output loading capacitors, C_3 through C_9, are mounted toward the rear so the leads from the tank coil can be kept as short as possible. A length of copper strip is used between the coil and the stator of C_3; originally this lead was No. 14 wire but on 10 meters the tank current was enough to heat it to the point of discoloration. The ground lead from the fixed units, made to the rear bearing connection of C_3, is also copper strip. C_3 and S_3 are operated through extension shafts, using Millen flexible couplings to simplify the alignment problem.

Underneath the chassis, each 811A grid is bypassed directly to the socket-mounting screw nearest the plate choke (right-hand side of the socket in Fig. 6-70). The d.c. leads have small chokes, RFC_2 and RFC_3, with additional bypasses for good r.f. filtering, particularly at v.h.f. since grid rectification generates harmonics in the TV bands. The filament choke, RFC_1, is mounted

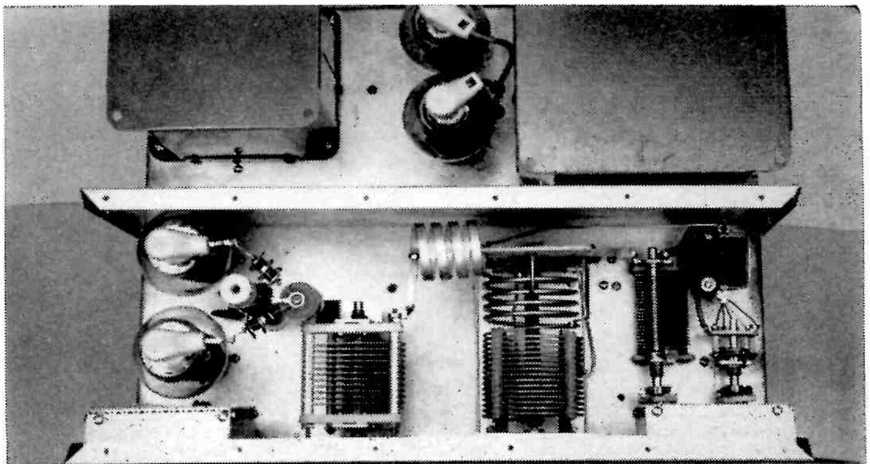


Fig. 6-69—The r.f. section with the shield cover removed. Components here are readily identifiable by reference to the circuit diagram. The meters are enclosed in rectangular boxes made from thin aluminum sheet, formed to be fastened by the meter mounting screws. The back covers on these boxes are made from perforated aluminum, folded over at the edges and held on the boxes by sheet-metal screws. The switch for shifting the 0-500 milliammeter (left) from grid to cathode is concealed by the box which encloses the meter.

so that the filament side is close to the filament terminals on the tube sockets; the other end is bypassed directly to the chassis.

The shielding around the amplifier consists of two pieces of sheet aluminum and a perforated aluminum ("do-it-yourself" type) cover having the shape of an inverted U. Fig. 6-69 shows how the rear wall is made. Its edges are bent to provide flanges for fastening the cover with sheet-metal screws, and there is a similar flange projecting to the rear at the bottom for fastening the wall to the chassis. The front piece extends the full height of the panel and is identically drilled and cut out for meters and controls. It has flanges at the top and extending down the sides from the top to the chassis. The cover itself extends down over the sides of the chassis for about one inch. Numerous screws are used for fastening the cover, to prevent leakage of harmonics.

The shields over the meters are made as described in the caption for the inside top view. Meter leads are bypassed to the shield boxes where they emerge.

Construction of the directional coupler parallels that given for the antenna coupler in Chapter Thirteen.

Operating Conditions and Tuning

The voltage delivered by the power supply is approximately 1500 volts with no drive and with the tubes taking only the no-bias static plate current, which is about 60 ma. At the full load of 350 ma. the voltage is slightly under 1400. Optimum operating conditions for 1400 volts at 350 ma. peak-envelope power input as an s.s.b. linear call for a peak-envelope grid current of 60 ma. The peak-envelope tube power output is close to 350 watts under these conditions. The same operating conditions are also about optimum for c.w.

The behavior of the cathode current when tuning a grounded-grid triode amplifier is somewhat confusing, and the meter is principally useful as a check on operating conditions rather than as a tuning indicator. The best indicator of proper tuning of the plate tank capacitor is the forward-power reading of the directional coupler. For any trial setting of the loading controls and driving power, *always* set the plate tank capacitor control at the point which results in a maximum reading on the power-output indicator.

The power indications are only relative, of course, and the sensitivity control should be set to give a reading in the upper half of the scale of the meter.

The objective in adjusting loading and drive is to arrive at maximum power output simultaneously with a plate current of 350 ma. and a grid current of 60 ma. — that is, a total cathode current of 410 ma. when the grid current reading is 60 ma. The loading is critical. If the amplifier is not loaded heavily enough the grid current will be too high and the right value of total cathode current either will not be reached or, if reached, the amplifier will be operating in the "flattening" region as an s.s.b. linear. (It can be operated this way on c.w., however, since linearity is unimportant here.) If the loading is too heavy, the grid current will be low when the cathode current reaches the proper value, but the efficiency will be low and the tubes will overheat.

Getting the knack of it takes a little practice, but when the job is done right the tubes will run cool on all bands in regular operation. Running key-down over a period of time may show just a trace of dark red color on the plates since the input and dissipation are somewhat over ratings under these operating conditions, although perfectly satisfactory with ordinary keying or s.s.b. voice.

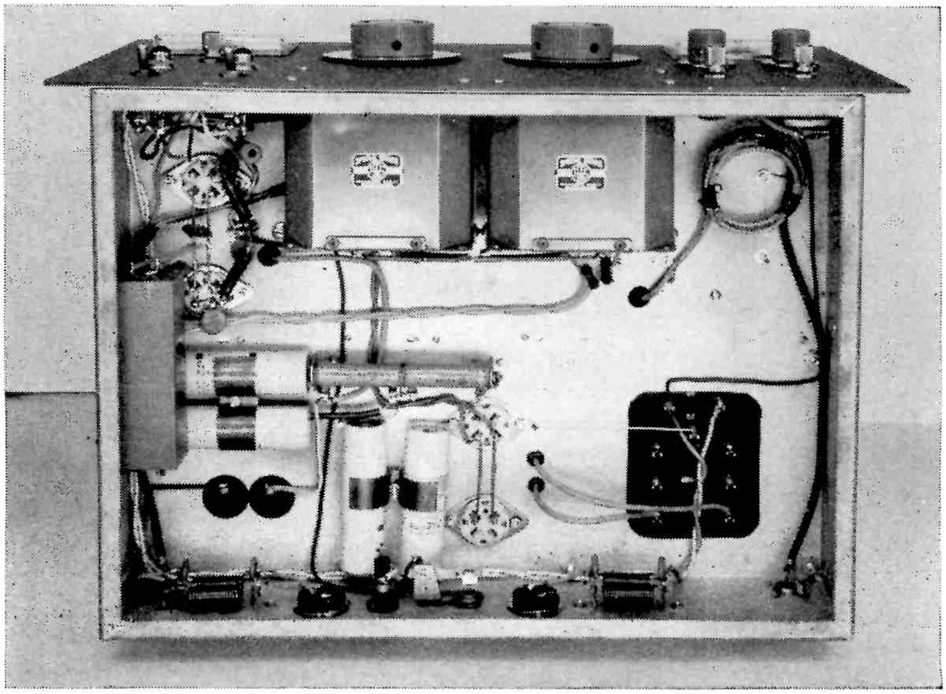


Fig. 6-70—In this below-chassis view, the two filament transformers are at the top, mounted on the chassis wall. The 811A sockets are at the upper left. The rectangular box on the left-hand wall contains the FC15 filament-choke assembly. The "Mickey Match" directional coupler is at the upper right. Filter capacitors and the bleeder resistors are in the lower section. A.c. inlets, fuse holder, bias jack, and the 115-volt line TVI filters are on the bottom chassis wall.

A Compact 650-Watt Amplifier

Compactness in the high-power amplifier shown in Figs. 6-71 through 6-76 is achieved through the use of germanium rectifiers in the power supply and tubes of the radial-beam type. When driven by an exciter delivering about 30 watts output, the amplifier runs at about 650 watts input and gives an output of about 400 watts on c.w. or p.e.p. s.s.b. It covers 80 through 10 meters by means of band switching and has a fixed 50-ohm output impedance.

Two 4X250B tubes operating Class AB_2 are used in a grounded-cathode circuit (see Fig. 6-72). No grid tuning is used, since an exciter of the size mentioned will drive the grids directly across the 110-ohm resistance. L_1 is a series peaking coil to increase the drive on 10 meters. A parallel-tuned tank with fixed-link output coupling is used in the plate circuit. This system has the advantage that series plate feed can be used, and no large output capacitance is needed. Tuning is straightforward and the coupling, once adjusted holds over a wide frequency range.

The link circuit is grounded through a removable jumper at the output connector, so that a balanced load can be fed if desired.

The small 15-$\mu\mu$f. capacitor (CRL Type 850), from the plates to ground, provides a short path for harmonic currents and keeps them out of the output coil. On the 3.5- to 4-Mc. range a fixed 100-$\mu\mu$f. capacitor is connected across the coil, so that a proper L-to-C ratio can be maintained at 4 Mc. When switched out of the circuit, the coil and fixed capacitor resonate around 5 Mc., which is sufficiently removed from any of the other ranges to avoid any difficulty.

The 10-ohm resistor in the B + lead serves as a fuse in case of a shorted tube or other fault that might endanger the power supply.

Power Supply

The plate supply uses two voltage doublers in series; see Fig. 6-75. Two 325-volt windings on T_2 feed strings of germanium rectifiers in full-wave voltage-doubler connections. Each doubler capacitance is 160 μf., made up of two parallel 80-μf. 450-volt cartridge type units with cardboard sleeves. The chassis is lined with insulating material under the C_5 and C_6 capacitors, since their outer cans run as high as +1300 volts. The ripple is around 3 per cent r.m.s., and the regulation from no load to full load is about 15 per cent. Sixteen cells are used. Each group of four cells in one side of a voltage doubler has two 560 K resistors connected across pairs of cells to equalize the reverse voltage drop. Other 560 K resistors are connected as bleeders only as a safety measure, since no bleeders are needed for proper circuit operation. But even with the bleeders, the capacitors can retain a charge for several minutes, so be careful!

Grid bias is furnished by a 75-volt winding on T_1, a half-wave rectifier and an 80-μf. capacitor. About −90 volts is developed across C_9 and applied to the tubes during stand-by periods. The operating bias is adjustable from −30 to −60 volts by R_3.

Screen voltage is taken from the +375-volt point of the plate supply (junction C_7 and C_8). It is dropped through the 6BF5 regulator to deliver a low-impedance output adjustable from about 250 to 325 volts at up to 75 ma. Since this type of regulator will not handle reverse current, bleeder R_2 (Fig. 6-72) is provided to offset no-signal negative screen current to the 4X250Bs and make the screen meter read on scale.

When in operating condition, the "reference" voltage for the screen regulator is the −90 volt bias supply. In stand-by condition the reference is switched down to the tap on R_3, thus reducing the screen voltage from its nominal +300 or so to a lower value. This action, together with the increased grid bias, insures that the 4X250Bs

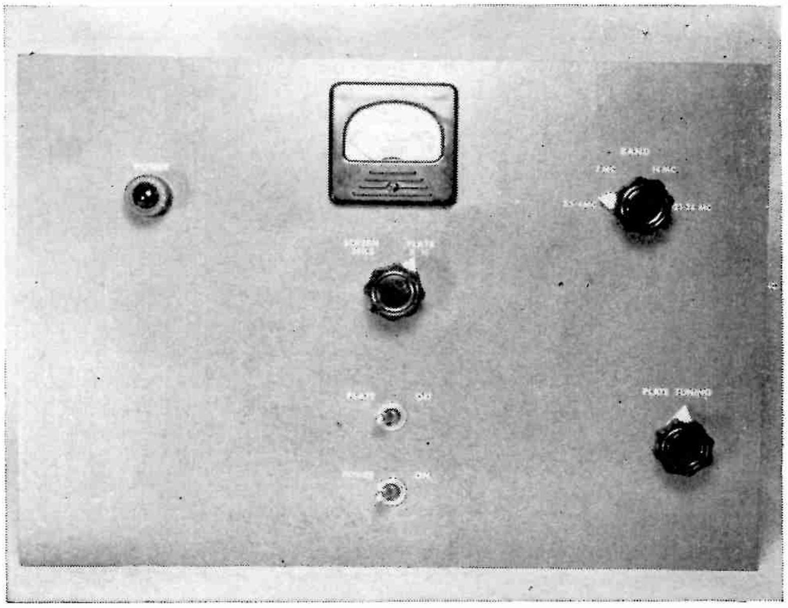


Fig. 6-71—The panel of this 650-watt amplifier built by W9LZY measures only 10 by 14 inches. Below the meter are the meter switch, high-voltage switch and filament/bias switch. To the right are controls for the band switch (above) and the tank capacitor (below).

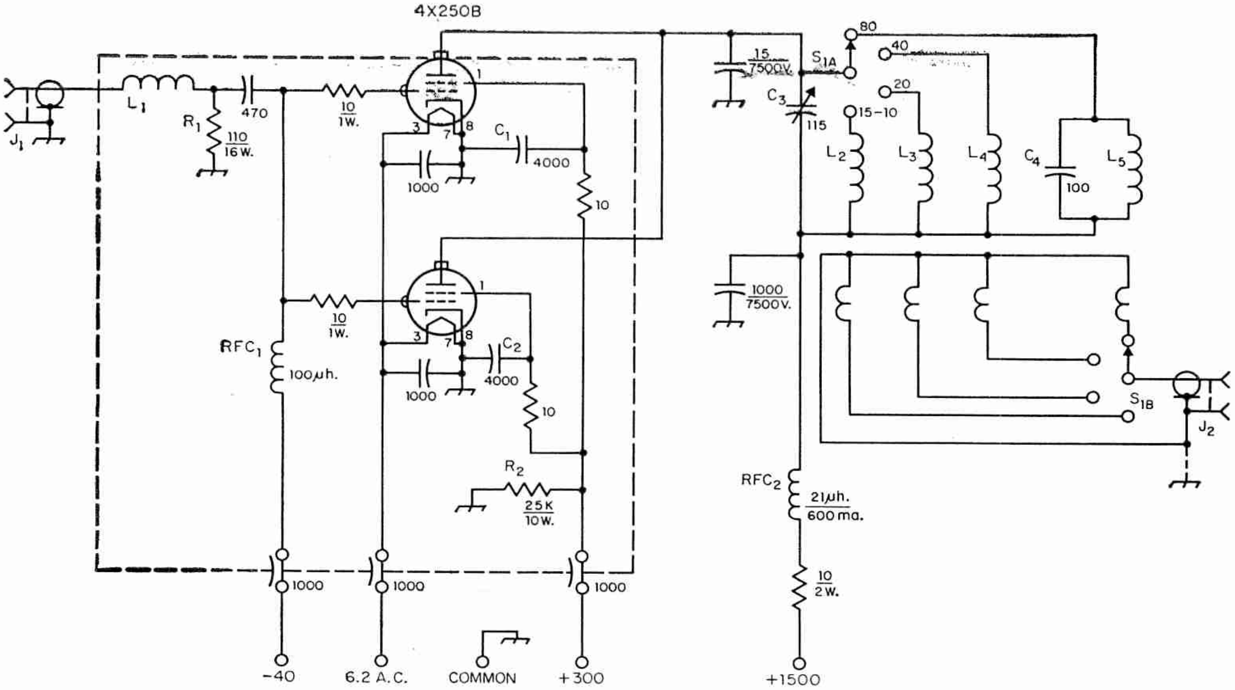


Fig. 6-72—Circuit diagram of the r.f. portion of the amplifier. Unless otherwise indicated, capacitances are in μμf., resistances are in ohms, resistors are ½ watt. The 1000-μμf. plate bypass is a CRL Type 858-S; the 1000-μμf. feed-through capacitors are 500-volt ceramic.

C1, C2—Four 1000-μμf. 500-volt disk ceramic capacitors in parallel.
C3—115-μμf. variable, 2000-volt spacing. See text.
C4—Two 25-μμf. NPO ceramic and one 50-μμf. N750 ceramic in parallel, 7500-volt rating.
J1—UG-291/U BNC panel jack (Amphenol 31-001).
J2—SO-239 UHF panel jack (Amphenol 83-1R).
L1—6 turns No. 20, ⅜-inch diam., ½ inch long.
L2—4½ turns ⅛-inch copper tubing, 1¼ inches long, 1⅛-inch diam. Link is 3 turns No. 16 wire, ¾ inch long, ¾-inch diam
L3—6 turns ⅛-inch copper tubing, 1½ inch long, 1⅝-inch diam. Link is 2 turns No. 12, ½ inch long, 1⅛-inch diam.
L4—8½ turns No. 12, 1⅛ inches long, 2⅛-inch diam. Link is 3 turns No. 12, ⅝ inch long, 1½-inch diam.
L5—Two coils, see text. Outer is 10 turns No. 12, 1⅜ inches long, 2⅛-inch diam. Inner coil is 6½ turns No. 12, ¾ inch long, 1¾-inch diam., inside plate end of outer coil. Link is 4 turns No. 12, ½ inch long, 1½-inch diam.
RFC1—100-μh. r.f. choke (National R-33-4).
RFC2—21-μh. 600-ma. r. f. choke (Ohmite Z-28).

draw no current in standby condition. In operation the grid, screen, and plate voltages all tend to vary in proportion to line-voltage changes.

The screen current is measured by switching the 0-75 milliammeter across 22 ohms in the lead to the screen-voltage regulator. The resistor has negligible shunting effect. For measuring plate current the meter is switched across a low resistance R_6, connected between the two sections of the plate supply. R_5 was adjusted for

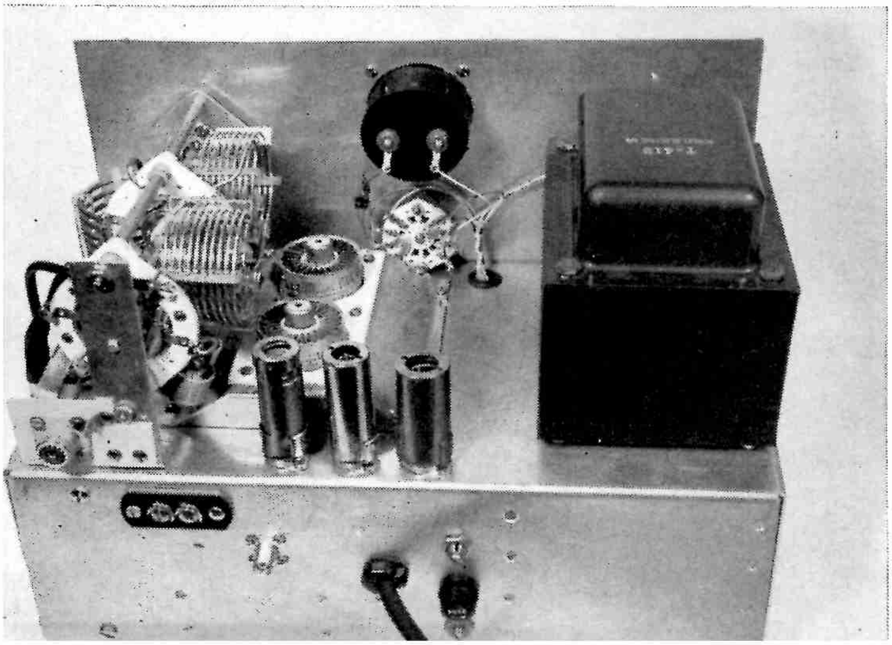


Fig. 6-73—Rear view of the 650-watt amplifier showing mounting of the 4X250Bs and the plate transformer. Shields in the foreground enclose voltage-regulator tubes and a relay. The shaft protruding from the rear edge of the chassis operates the bias potentiometer.

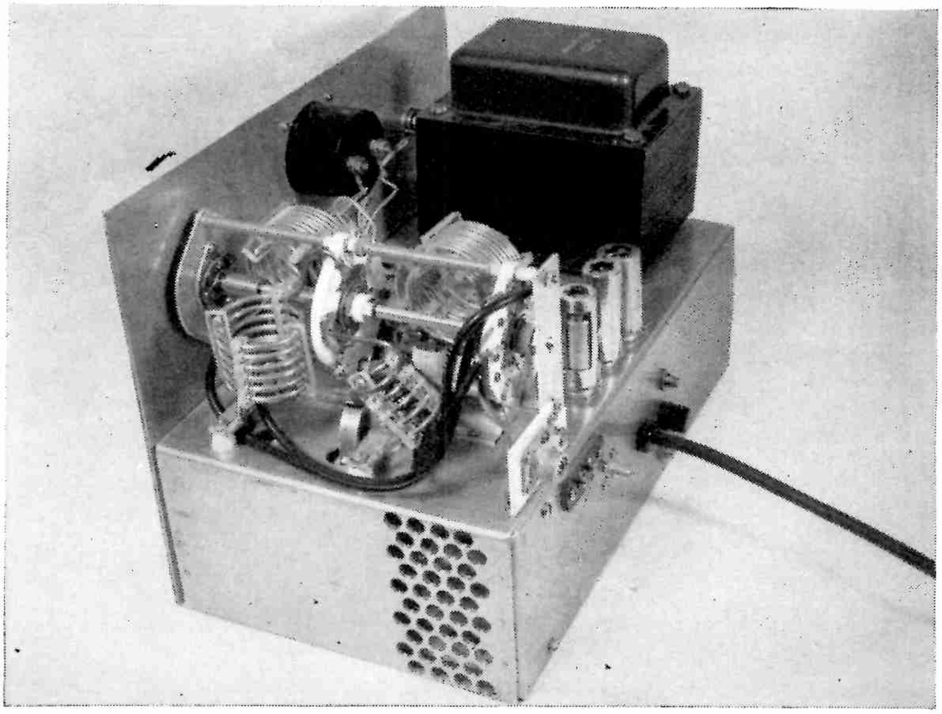


Fig. 6-74—Side view of the 4X250B amplifier showing mounting of the band switch and tank coils. The chassis is perforated for ventilation.

full-scale meter reading at 750 ma. There is a maximum of 425 volts between switch contacts and 850 volts from contacts to ground.

The stand-by relay K_1 is one that plugs into a 7-pin miniature socket. It operates from 115 volts a.c. and a half-wave power supply. The input is brought out to two terminals on the rear of the chassis, where connection is made across the antenna relay coil.

Construction

The amplifier is built on an 8 × 14-inch chassis with a 10 × 14-inch panel. The chassis is 4½ inches deep, to provide space for the filter capacitors and cooling fan underneath. As can be seen by studying the photographs, the plate power supply occupies the left end of the chassis, and the r.f. circuits take most of the remaining space. The heater and bias supply is stowed under the right rear corner of the chassis behind the plate tuning capacitor. The screen regulator and stand-by relay are at the rear of the chassis in the center.

The controls are few and simple. The band switch has four positions, for the 80-, 40-, 20- and 15- and 10-meter bands. Other controls are the plate tuning capacitor, plate-current/screen-current meter switch, power and plate voltage switches.

The plate tank capacitor is one from a BC-375 tuning unit, mounted under the chassis on four ceramic feed-through bushings. (Any other capacitor of equivalent rating, such as the Johnson 155-4 may be substituted.) Four holes were drilled and tapped in the ¼-inch square frame rods on the right-hand side of the capacitor, and 6-32 threaded rod was screwed into the holes and passed through the insulators. The four screws project above the insulators at the top of the chassis, where the B+ ends of the plate coils connect to them via copper strips. An insulated shaft extension goes through the panel to the tuning knob.

The wire from each coil was wrapped around a pipe of suitable diameter. Four Plexiglass strips were drilled with clearance holes at the desired spacing, then the coil wire was fed through the holes. The 80-meter coil was made with two concentric sections in series to get enough inductance into the available space. The 80- and 40-meter links were also threaded through strips, while the 20- and 10-meter links are self-supporting. All links are a push fit inside the insulating strips of their respective coils, and are held with a drop or two of cement after adjustment.

The two band-switch wafers are each single-pole, 4-position, 60-degree throw (Communications Products Co., Type 86). A 60-degree index-and-shaft assembly from an Oak Type H switch was used. The rest of the switch was made up from 6-32 threaded brass rod, ¼-inch o.d. tubing, 1/16-inch aluminum sheet, and miscellaneous ceramic spacers and fiber washers from junked rotary switches.

The front wafer switches the plate coils. The links are connected to the rear wafer through RG-58/U cable, except the 80-meter link which goes direct. The cold sides of all links are soldered to a strip of copper running around the wafer, supported by 2-56 screws through the unused holes between contacts. The u.h.f.-type output connector is mounted on a strip of bakelite fastened to the rear switch bracket; its shell is grounded through a couple of solder lugs shown. T_2 weighs about twenty pounds; the chassis should be at least 0.08-inch aluminum to be strong enough to carry it.

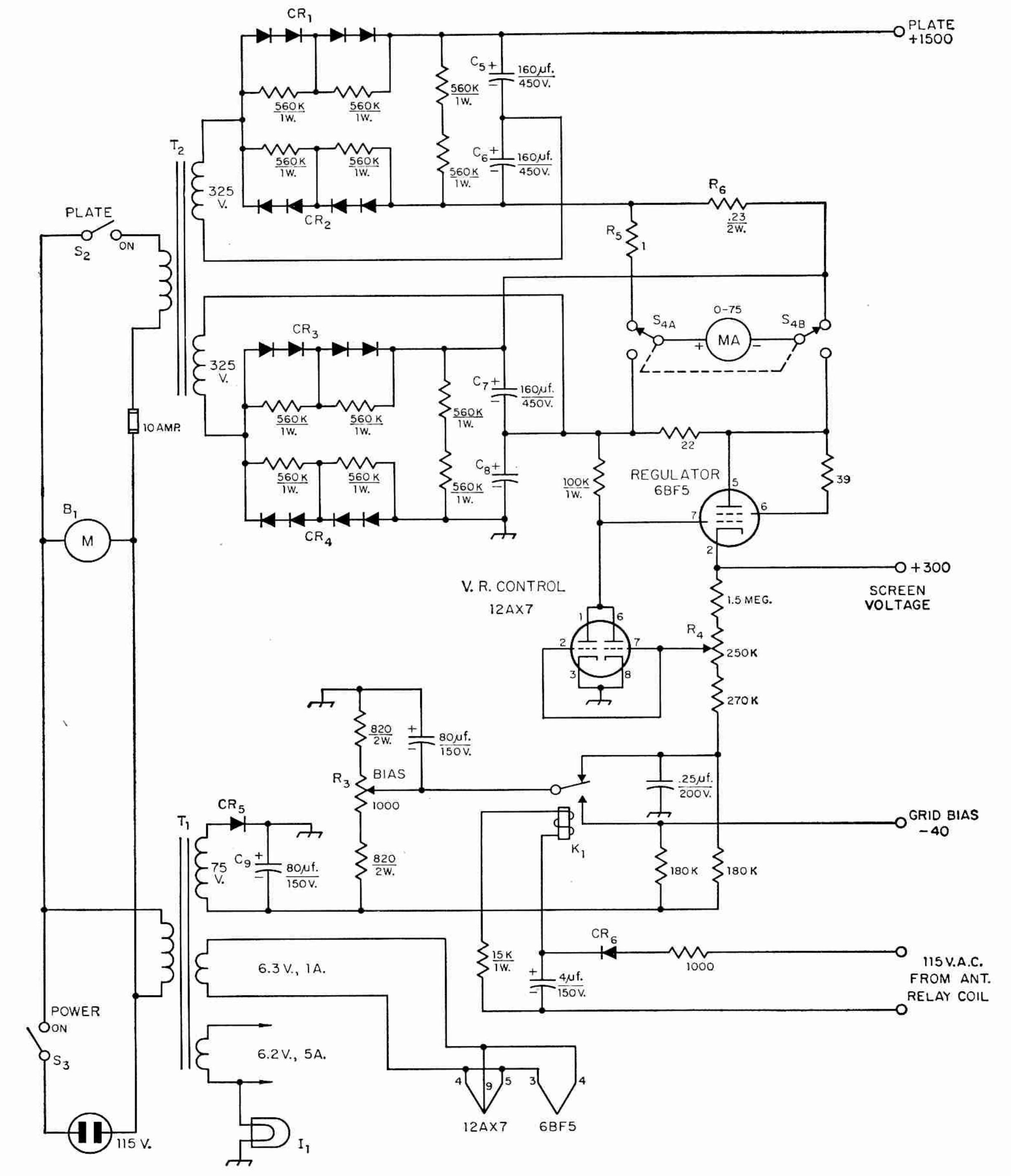


Fig. 6-75—Circuit of the power supply. Unless otherwise indicated, resistances are in ohms, resistors are ½ watt.

B_1—3250-r.p.m. motor with 4-inch fan blade (Rotron* 92-AS motor).

C_5 C_6, C_7, C_8—Two 80-μf. electrolytics in parallel (Sprague TVA-1716). Insulate as described in text.

CR_1—CR_4—Four 500-ma. 300-volt peak inverse (1N153 or equiv.).

CR_5—100-ma. 380-volt peak inverse.

CR_6—65-ma. 380-volt peak inverse (Federal 1002A).

I_1—150-ma. 6-8 volts (GE No. 47).

K_1—5000-ohm coil, 4 ma. pull-in (Terado Series 600 or equivalent).

R_2—2-watt linear potentiometer (Ohmite CU-1021).

R_4—2-watt linear potentiometer (Ohmite CU-2541).

S_2, S_3—15-amp. 125-volt toggle (Cutler-Hammer 7501-K13).

S_4—Two-pole 2-throw 60-degree throw ceramic rotary switch, non-shorting. See text.

T_1—6.2 volts at 5.5 amp., 6.3 volts at 1 amp., 75 volts at 100 ma. (Forest Electric Co.** T-423).

T_2—Two-secondaries, each 325 volts, 1 amp. (Forest** T-412).

*Rotron Mfg. Co., 7 Schoonmaker Lane, Woodstock, New York.

**Forest Electric Co., 7216 Circle Rd., Forest Park, Ill.

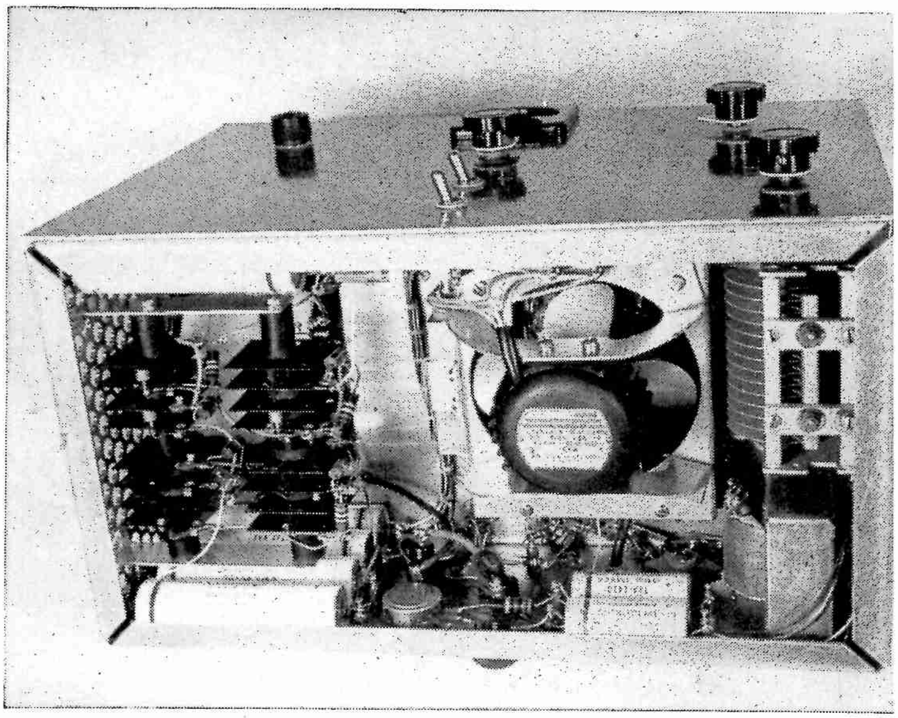


Fig. 6-76—This bottom view show the ventilating fan, tank capacitor, rectifier stacks and filter capacitors.

A bottom cover and a perforated-metal shield over the top, sides and rear should be added, for safety as well as TVI-proofing. An opening should be cut above the r.f. tubes and covered with hardware cloth.

Cooling

Each 4X250B tube requires at least 3.6 cubic feet of air per minute through the anode cooler. The base also requires some air. The tube is ordinarily mounted in an Eimac "air-system" socket so that the air flows first over the base, then through the anode cooler. This leads to a fairly large pressure drop, which is ordinarily considered to require a centrifugal blower. Since a blower of this type requires considerable space, the design has been altered to permit the use of a fan. Only the insulating rings and contacts from Eimac sockets are used, mounted by the cathode tabs in oversized holes in the chassis. Many small holes are drilled in the chassis to provide additional air passage. A small aluminum housing above the chassis directs all the air through the anode coolers. It comes to within ¼ inch of the anode coolers. The opening is closed by a piece of Fiberglas-base plastic fitting on top. It comes to within 1/16 inch of the tubes, so that a small amount of the air flows around the outside of the coolers.

All of the left end and part of the right end of the chassis are perforated by ⅜-inch holes. The air drawn in by the fan passes over the plate rectifier fins and past the heater transformer. The whole air path is direct and free from large obstructions and sharp bends.

The fan is a 4-inch blade driven by a Rotron Mfg. Co. Type 92-AS motor at 3250 r.p.m. It is mounted in a hole 4⅛ inches in diameter in the grid housing, with about ⅓ of the blade thickness projecting into the housing. The motor is a capacitor-run type. The 1-μf. 600-volt phasing capacitor mounts on the side of the grid housing. The motor, housing and capacitor can be removed as a unit, leaving only the front and rear walls of the housing in place.

Under the conditions described, the pressure *vs.* flow curves of the fan and of the tubes indicate that somewhere around 10 c.f.m. of air is delivered. This is entirely ample for the pair of 4X250Bs. Since the only major source of heat is the tubes, and since this heat is quickly removed by the air, the whole amplifier runs at a satisfactorily low temperature.

Operation

For Class AB_2 operation, the screen voltage is set at 300 volts, and the grid bias at a point (about −40 volts) where the tubes draw 150 ma. without drive. When operating and fully loaded, full output from an HT-30 or similar exciter should swing the plate current to approximately 400 ma.

The various links are of approximately the right inductance to couple to a 50-ohm load. They must be quite tightly coupled to their plate coils. When properly positioned with a 50-ohm load connected, the plate current dips 10 or 15 ma. as the plate capacitor is tuned through resonance with r.f. drive applied. Once adjusted, these links are left alone. The antenna is tuned with the aid of an s.w.r. bridge to present a 50-ohm load to the amplifier. The amplifier should not be operated without a suitable load

Operation is now very simple. The heaters are warmed up for at least 30 seconds. With the plate power switch *off*, the band switch is set to the proper range. The exciter is tuned up to give c.w. output. (Not more than 40 volts r.m.s.) The plate power is turned on and the plate capacitor tuned to the plate current dip, or to maximum indicated output if a Micromatch is being used. The exciter is then set to give the type of output desired.

(Originally described in *QST* for Sept. 1958.)

4-250-A's in a 1-Kw. Final

The amplifier shown in the accompanying photographs uses two 4-250As in parallel and covers 3.5 to 28 Mc. with complete band-switching. The output circuit is a pi network designed for working into reasonably well-matched 52- to 75-ohm coaxial lines. The amplifier can handle a kilowatt input in Class C operation on either phone or c.w. without pushing the tubes to their limits. It can also be operated as a linear amplifier for single side band.

The various components are mounted on a 17 × 13 × 4-inch aluminum chassis attached to a standard 19-inch relay rack panel 15¾-inches high. The above-chassis section is enclosed in a 11½-inch high shield made from 1/16-inch sheet aluminum. An aluminum bottom plate completes the below-chassis shielding. Enclosing the amplifier in this way, plus the use of shielded wire and filters in the supply leads, takes care of the harmonic TVI question.

The 4-250As are cooled by forcing air into the chassis and thence up past the tubes by means of a 21 cu. ft. per minute blower. The air is exhausted through two 3-inch diameter circular openings over the tubes in the top cover. To maintain the shielding intact, these are covered with perforated aluminum.

A Barker and Williamson Model 850 band-switching pi-tank inductor is used in the output circuit. It is tuned by a vacuum variable capacitor operated through the counter dial (Groth TC-3) shown in the panel view

Circuit Details

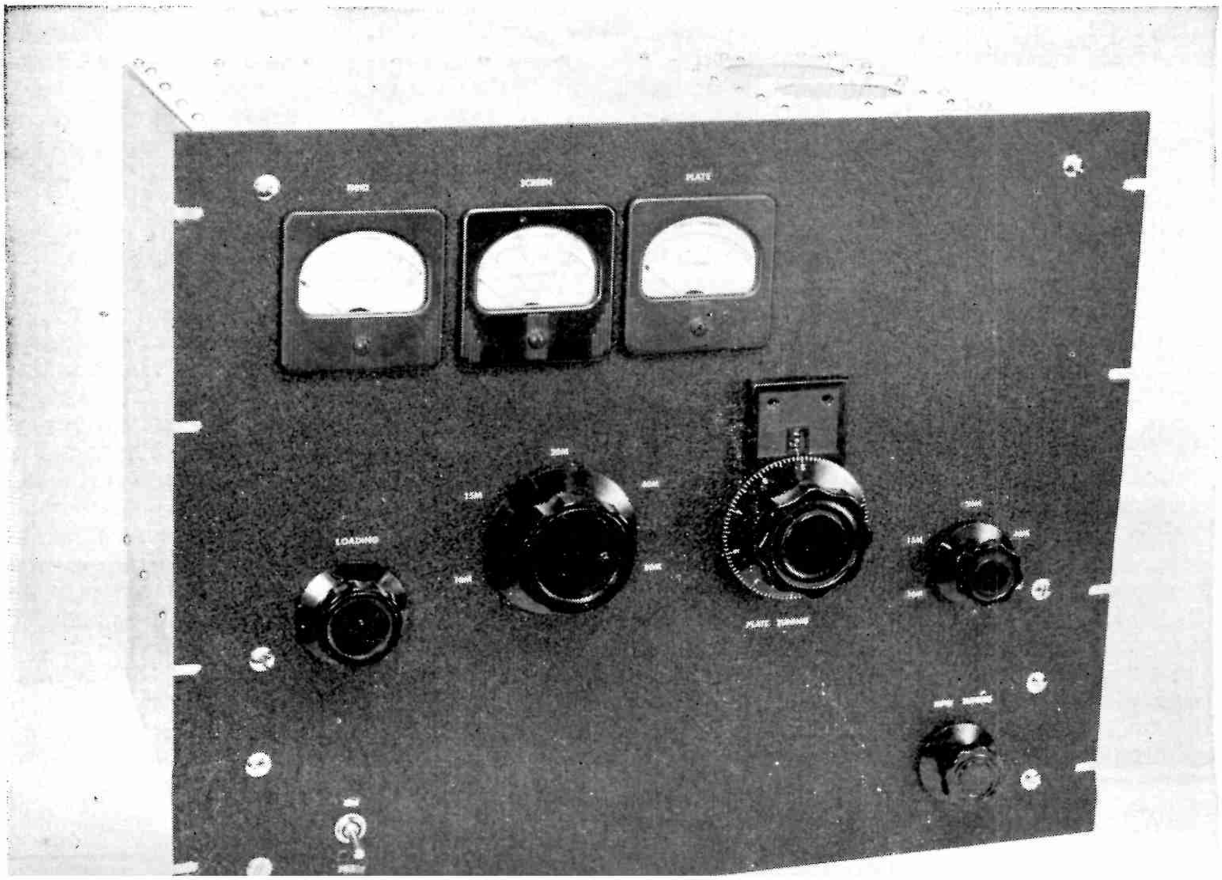
The circuit, Fig. 6-78, is electrically the more-or-less standard arrangement of a parallel-tuned grid circuit and a pi-network output circuit. The amplifier is neutralized by the capacitive bridge method. A filament transformer is included, but all other voltages come from external supplies.

The grid input circuit of the amplifier uses a slightly modified B&W turret assembly. The grid coils are tuned by a 75-µµf. variable. The 20-, 15-, and 10-meter coils each must have a few turns removed for proper grid tuning on these bands.

The circuit includes a 2000-ohm grid leak and has provisions for external bias, which should be used in combination with the leak. The bypass capacitors on the screen leads all carry a rating of 1600 volts. This rating is necessary to avoid capacitor breakdowns when operating the amplifier screens at their rated voltages for AB_1 operation, and also with plate-modulated Class C operation where the 600-volt rating of the smaller ceramic capacitors would be exceeded on modulation peaks. All of the 0.001- and 0.003-µf. capacitors are the disk type, and aside from the screen bypasses are used mainly for filtering TV harmonics from the supply leads.

The bypass capacitors in the high-voltage lead

Fig. 6-77—A 1-kw. final using a pair of 4-250-A's in parallel.



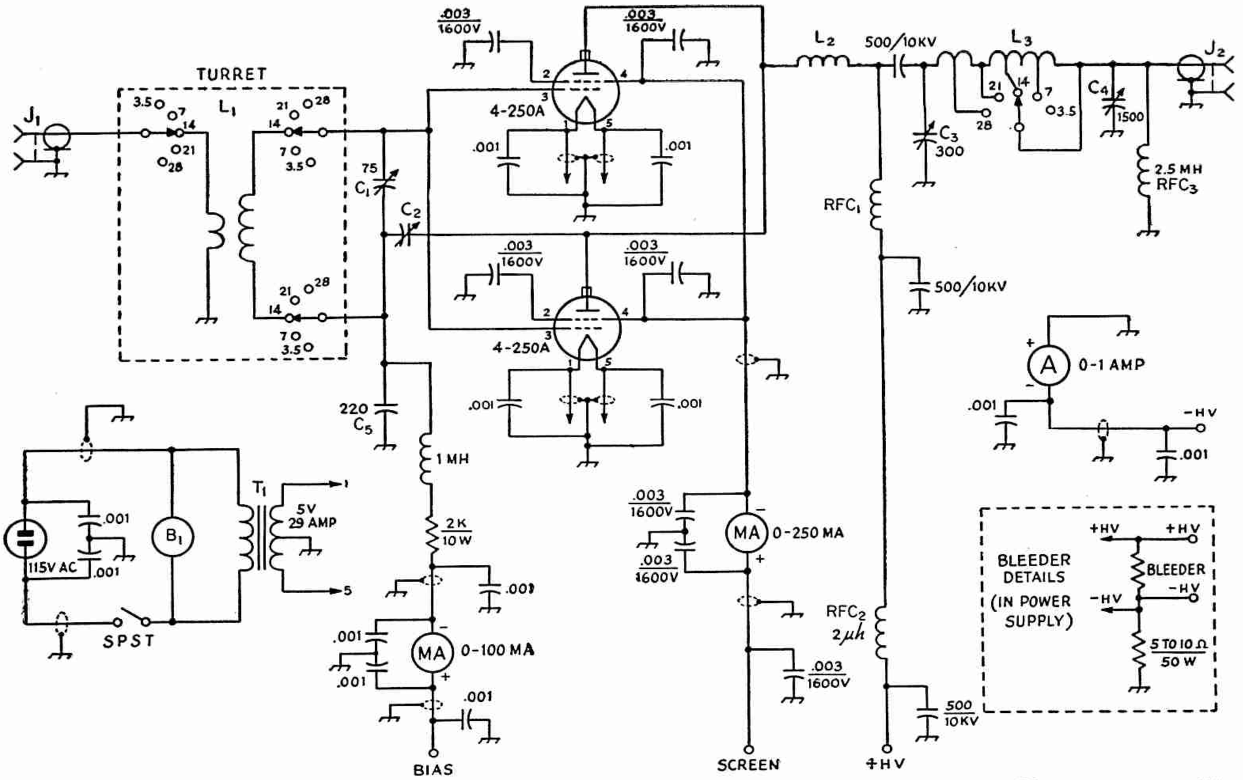


Fig. 6-78—Circuit diagram of the 4-250A amplifier.

B1—Blower-motor assembly, 21 c.f.m. (Ripley model 8433).

C1—75-μμf. variable, receiving spacing (Millen 19075).

C2—7-μμf. neutralizing capacitor (Cardwell type ADN).

C3—300-μμf. vacuum variable (Jennings type UCS).

C4—1500-μμf. variable (Cardwell type 8013).

C5—220-μμf. mica or NPO ceramic.

J1, J2—Coax receptacle, chassis mounting.

L1—Turret assembly (B&W BTEL with 14-, 21-, and 28-Mc. coils modified by removing turns).
3.5 Mc.: 39 turns No. 22, 1¼ inches diam., 1⅜ inches long, link 3 turns No. 18.
7 Mc.: 20 turns No. 20, 1¼ inches diam., 1W inches long, link 3 turns No. 18.
14 Mc.: 8 turns No. 18, 1¼ inches diam., ¾ inch long, link 2 turns No. 18.
21 Mc.: 4 turns No. 16, 1¼ inches diam., ½ inch long, link 1 turn No. 18.
28 Mc.: 2½ turns No. 16, 1¼ inches diam., ½ inch long, link 1 turn No. 18.

L2—V.h.f. parasitic suppressor, 4 turns No. 12, ¼ inch dia., turns spaced wire diameter.

L3—Pi-tank inductor (B&W Model 850). Inductances as follows: 3.5 Mc., 13.5 μh.; 7 Mc., 6.5 μh.; 14 Mc. 1.75 μ.; 21 Mc., 1 μh.; 28 Mc., 0.8 μh.

RFC1—National type R175A r.f. choke.

RFC2—2-μph. 500-ma. r.f. choke (National type R-60).

RFC3—2.5-mh. r.f. choke.

T1—Filament transformer, 5 volts, 29 amp. (Thordarson T-21FO7-A).

are the TV high-voltage ceramic type, as is also the blocking capacitor in the tank circuit. The loading capacitor, C_4, in the output circuit of the amplifier is a variable having enough range (1500 μμf. total capacitance) to give adequate loading on 80 through 10 meters when working into a 52- or 75-ohm resistive load.

Plate current is metered by a 0–1 ammeter shunted across a resistor in the negative high-voltage lead. As shown in Fig. 6-78, this resistor is incorporated in the power supply, not in the amplifier unit. A 50-watt rating represents an ample safety factor, since the power dissipated would not exceed a few watts should the ammeter open up.

Separate milliammeters are provided for the grid and screen circuits. The screen meter is quite essential since the screen current, and hence screen dissipation, is very sensitive to grid driving voltage and plate tuning.

Layout Details

Fig. 6-79 is a view looking into the amplifier with the top cover removed. The variable capacitor at the right is the output loading control, C_4. To the left of C_4 is the Model 850 inductor unit. Immediately to the rear (below, in the photograph) of the inductor is the output lead, connected to a coaxial receptacle mounted on the rear cover. The vacuum variable, C_3, is mounted between the inductor and the 4-250As. It is supported by an aluminum bracket 6 inches high and 4 inches wide. The neutralizing capacitor C_2 is between the 4-250As and the front panel.

The grid turret and tuning capacitor are mounted underneath the chassis to take advantage of the shielding afforded thereby. To fit under the chassis the turret is mounted with the switch shaft vertical, necessitating a right-angle drive to the panel control. The shaft approaches the panel at an angle, so a flexible coupling of the ball type (Millen 39001) is used between the shaft and panel bearing.

The meters are in a separate enclosure measuring 11 × 3 × 3-inches. It is mounted to the front of the box by countersunk flat-head screws. The top lips of the meter box are drilled to take sheet-metal screws when the lid is in place.

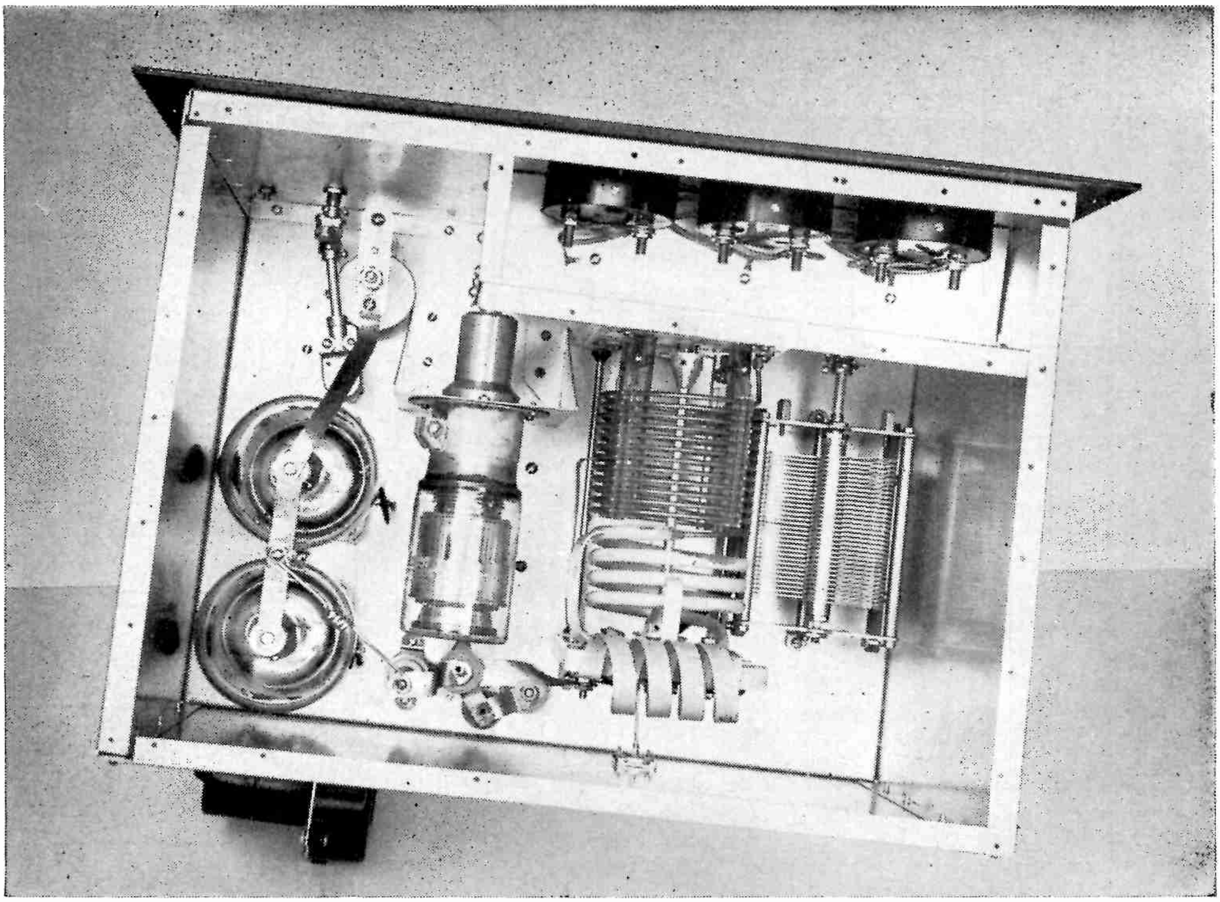
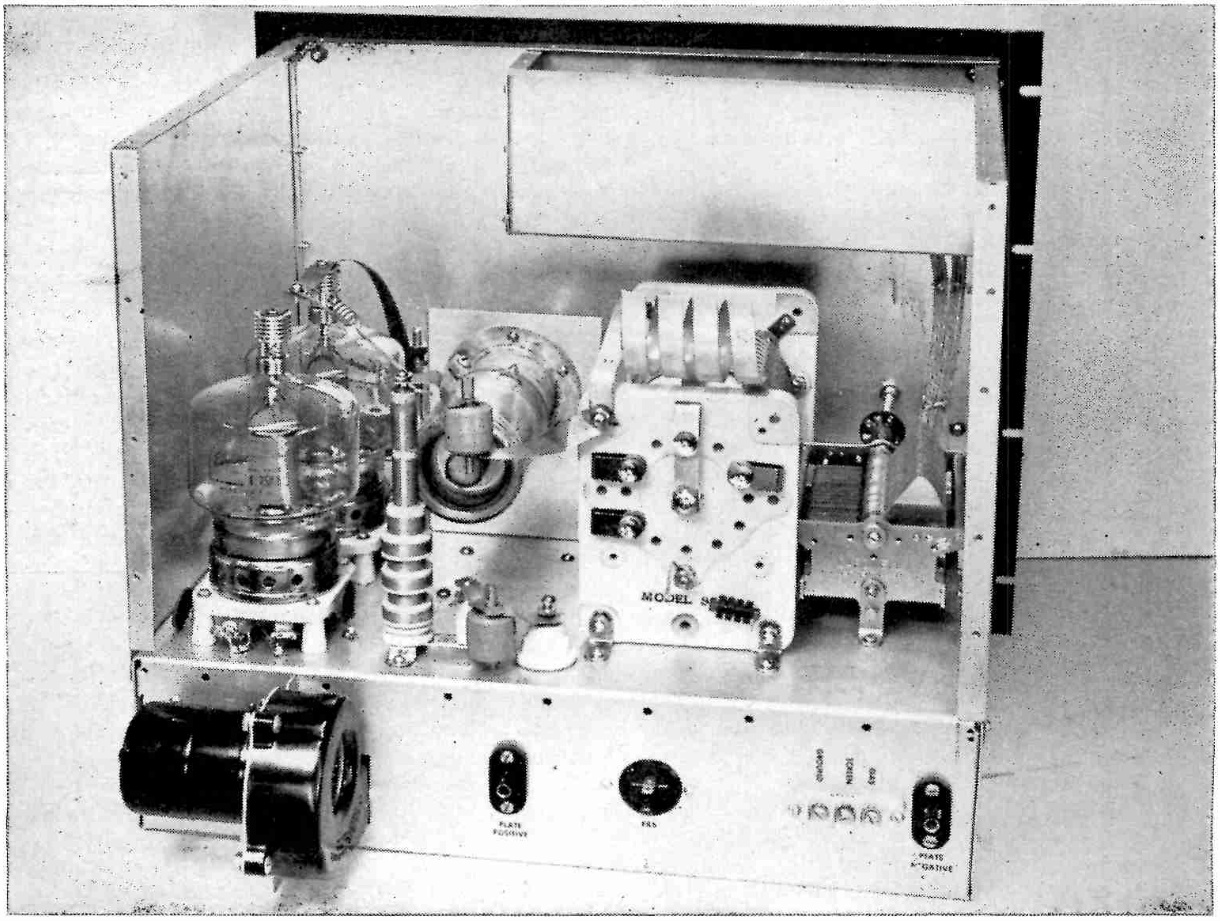


Fig 6-79 (above)

Fig. 6-80 (below)



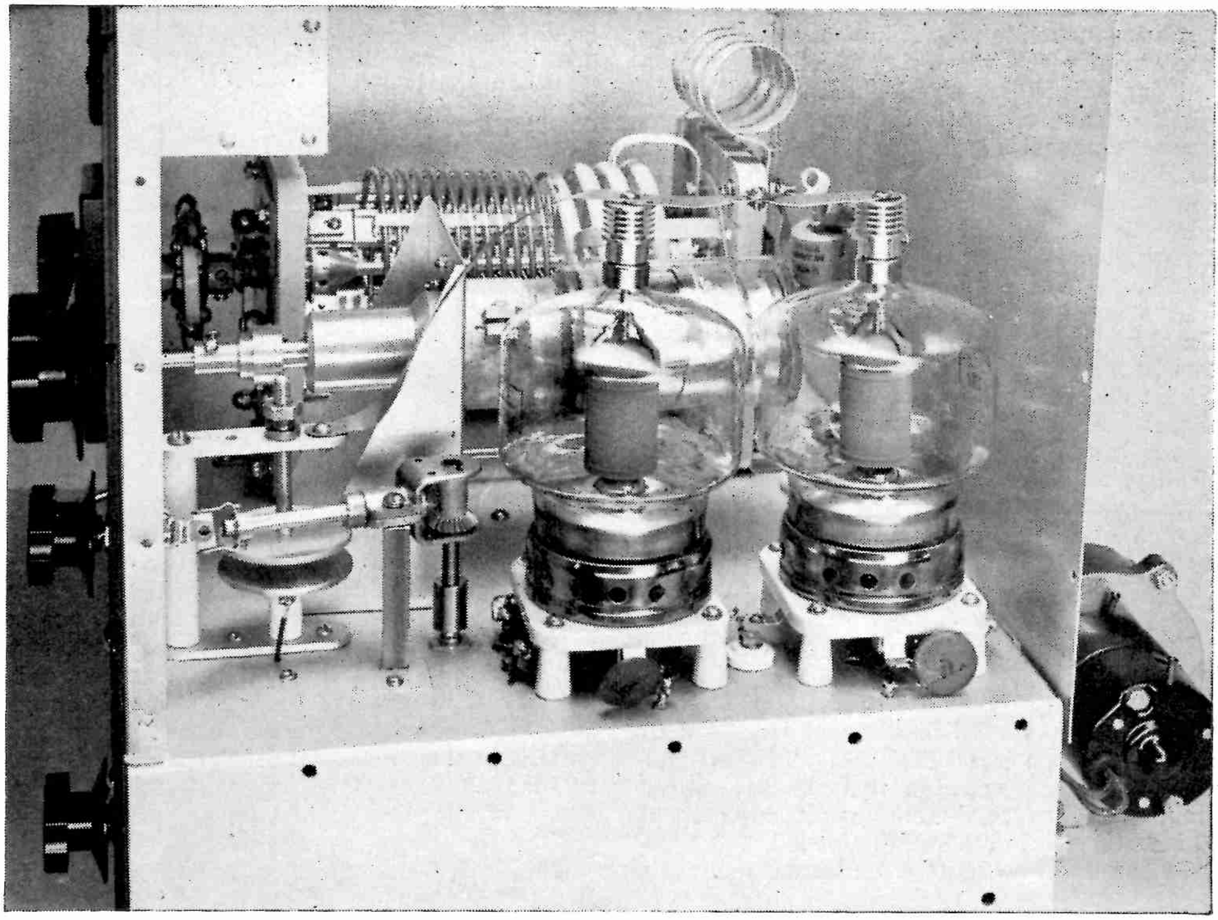


Fig. 6-81

Connections to the tube plates and neutralizing capacitor are made from flexible brass strip ½ inch wide. A piece of ¾-inch wide brass strip is used for the connection between the stator terminal of the vacuum variable and the tank inductor. The blocking capacitor is mounted on this strip.

Fig. 6-80 shows the amplifier with the top and back panels removed. The blower assembly is mounted on the rear chassis wall. To the right of the motor is the high-voltage terminal, the 115-volt connector, the grid and screen terminals, and the high-voltage negative connector. Leads from these last three terminals run below chassis in shielded wire and then up to the meter box. These leads are visible in front of the loading capacitor. Belden 8885 shielded wire is used for the leads. The inner conductor is bypassed to the shield braid at each end. The 2.5-mh. "safety" choke, *RFC*$_3$, shunting the output end of the pi network is mounted on the back of the tank coil between the output lead and chassis ground.

The isolantite feedthrough insulator to the left of the inductor is used to bring the high voltage through the chassis. Adjacent to it is the bypass at the bottom of the plate choke, *RFC*$_1$.

Mounting details of the right-angle drive assembly for switching the grid circuit are clearly visible in Fig. 6-81. A ½-inch square rod 2¾ inches long is drilled and tapped at both ends to support the drive.

The sockets for the 4-250As are mounted on one-inch isolantite pillars. The screen and filament terminals are bypassed directly at the socket terminals. The grid terminals on the sockets face each other, and a small feedthrough is used to bring the grid lead up through the chassis.

Fig. 6-82 is a bottom view of the amplifier and Fig. 6-83 is a close-up view of the grid circuit. A short length of RG-58/U is used to connect J_1 on the rear chassis wall to the link terminals on the turret assembly. The high-voltage lead is filtered by the 500-$\mu\mu$f. ceramic bypass and *RFC*$_2$. These two components are visible on the inside of the rear wall above the blower assembly. Two-terminal tie-points are used for the a.c. connections to the filament transformer and blower motor. Shielded leads are used between the tie-points and the 115-volt connector.

Fig. 6-83 shows the grid-circuit wiring in a bit more detail, particularly the grid choke, grid resistor and C_5 clustered just above the tuning capacitor. The modifications to the 10- and 15-meter coils also are somewhat more easily seen in this photograph.

Adjustment and Operating Data

The amplifier should be neutralized with the plate and screen supply leads disconnected and the bandswitch set to 28 Mc. An indicating wave meter should be coupled to the tank circuit and drive applied to the amplifier. Resonate the grid and plate tanks and adjust the neutralizing capacitor for minimum r.f. in the tank circuit as indicated by the wave meter. The same neutralizing adjustment should hold for all bands. Don't attempt to neutralize with the plate and screen supply leads connected — i.e., with a complete circuit for d.c. — because even with the power turned off this permits electrons to flow from the cathode to the plate and screen, and r.f. will be present that cannot be neutralized out.

The parasitic choke will, in general, resonate the plate lead in one of the low v.h.f. TV channels, and will tend to increase harmonic output in that channel. Measure the resonant frequency of the plate lead at L_2 with a grid-dip meter, and if it is in one of the channels received in your locality, either pull the turns apart, or squeeze them together to move the frequency to an unused channel. Any frequency from 70 to 100 Mc. should be satisfactory.

Power Supply

For 1 kw. input, a plate voltage of at least 2000 is required. Screen voltage is obtained preferably from a separate 400-volt supply. For Class C operation, an external bias supply regulated by a VR-150, plus a grid leak of 2000 ohms is recommended. With this combination, the grid current should be 25 ma. Screen current should be about 60 ma. with the amplifier fully loaded.

Some sort of r.f. output indicator, such as a crystal-rectifier voltmeter or r.f. ammeter in the feed line, should be used in tuning. It is preferable to do the preliminary tuning with the plate voltage applied to the tubes but with the screen voltage at zero. Zero screen voltage, provided the d.c. screen circuit is complete, will give enough output for tuning adjustments. C_2 and C_4 are adjusted to give maximum output, and the screen voltage is then increased until the amplifier is running at the desired input. C_3 is of course tuned for the plate-current dip so that the amplifier tank is kept tuned to resonance.

The fixed values of inductance available in the B&W unit preclude the possibility of matching over a wide range of impedances. The circuit can handle an s.w.r. in the coax line of about 2 to 1, but with higher s.w.r. values it may not be possible to get the desired loading. Also, although the construction is such that the amplifier is "clean" insofar as direct radiation and leakage of harmonics in the TV bands are concerned, a good low-pass filter will be required in most installations. A low s.w.r. in the coax line is definitely a requirement if excessive build-up of currents or voltages in the filter is to be avoided. If the line cannot be matched at the antenna, an auxiliary antenna coupler will have to be used.

For plate modulation a choke coil may be connected in the d.c. screen lead so the screen voltage will follow the audio variations in plate voltage. The choke should have an inductance of about 10 henrys, and must be capable of carrying 125 ma. d.c. For Class AB_1 operation on single side band the circuit may be left intact, the only requirement being to supply the proper operating voltages from suitably well-regulated supplies. If the amplifier is to be operated in AB_2 on s.s.b. the grid-leak resistor should be shorted out; also, suitable loading should be applied to the grid tank to maintain good regulation of the r.f. driving voltage.

(From *QST*, June, 1956.)

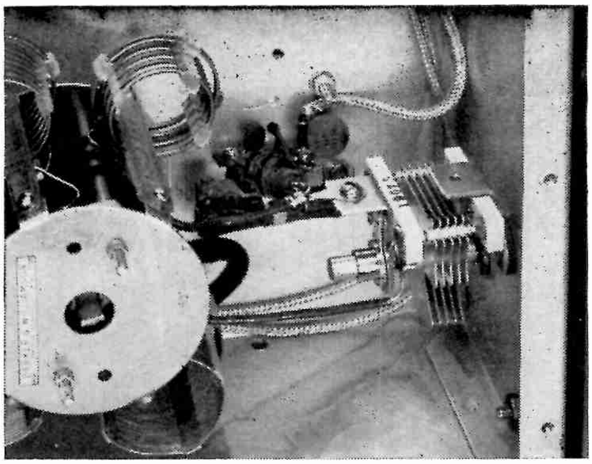
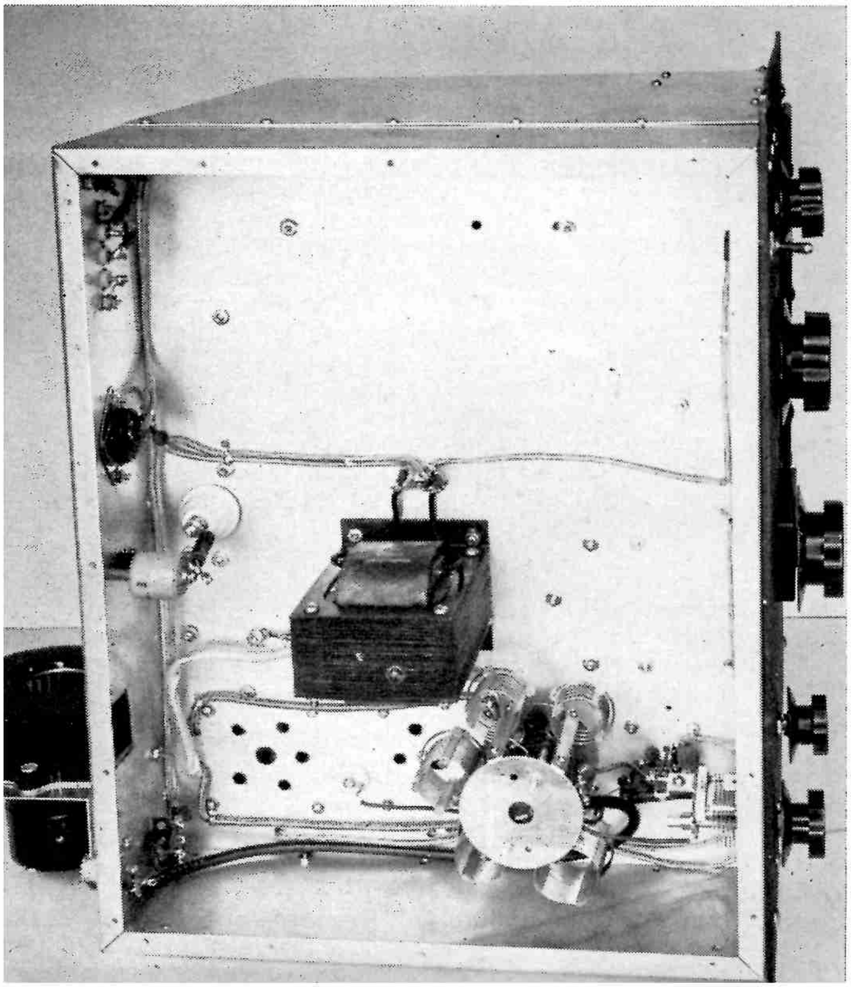


Fig. 6-83

Fig. 6-82


A V.F.O. With Differential Keyer

Figs. 6-84 through 6-88 show a v.f.o. with output on either 3.5 or 7 Mc. Included is a differential system for keying the control grid of an amplifier. The diagram is shown in Fig. 6-86. One section of a 12AT7 is used in the Vackar oscillator circuit, while the second section is used as a cathode follower driving a 5763 amplifier/doubler. S_1 selects either of two frequency ranges — 3.5 to 4 Mc. for use in the 80-meter band, and 3.5 to 3.65 Mc. for multiplying to the higher-frequency bands. If only the first range is desired, C_1 and C_3 may be omitted and the stators of C_2 and C_4 connected to the junction of C_5 and L_1. If both 3.5- and 7-Mc. output is desired, the two coils can be put on a switch section ganged to S_1.

To avoid chirp and permit full break-in c.w. operation, a differential keying system is used. Grid-block keying of an amplifier stage beyond the v.f.o. unit is provided by the negative power supply (6X5 rectifier), the 470K resistor, the 33K resistor R_1, and the 0.1-μf. capacitor C_6. The 6J5 cathode follower and the 0A2 control the oscillator. A complete description of the circuit operation will be found in Chapter Eight. Opening S_2 turns on the oscillator for "frequency spotting" purposes.

Construction

The unit is built on a 7 × 12 × 2-inch aluminum chassis that will fit inside an 8 × 14½ × 8¼-inch cabinet (Bud C-1747). The panel is 8 by 12 inches and the dial is a Millen 10035. Before mounting the components, it is advisable to stiffen the chassis against vibration by fastening two lengths of aluminum angle stock running lengthwise against the under surface of the chassis. Several machine screws should be used with each.

The v.f.o. tuned-circuit components are enclosed in a 4 × 5 × 6-inch aluminum box. This should also be stiffened with lengths of angle stock, one strip running under the top of the box, and one externally along each of the side covers.

The coil is supported on 2½-inch ceramic pillars (Millen 31002). The tuning capacitor C_4 is elevated above the bottom of the box on an aluminum bracket so that its shaft will line up with the dial. The band spread switch S_1 is mounted in the bottom of the box, to the rear of the coil, with its shaft vertical. The shaft is controlled from the panel by means of a National RAD right-angle drive and a "universal-joint" type shaft coupler (Millen 39001), as shown in the bottom-view photograph.

The three trimmer capacitors are mounted in the top of the box. C_3 is submounted so that its shaft, which is at high r.f. potential, will not protrude from the box. It is adjusted with an insulated screwdriver through a hole in the top of the box. C_5 is an air trimmer used here as a fixed capacitor. It is mounted on a bracket fastened to the bottom of the box, under the coil, and set at maximum capacitance.

The box should be placed on the chassis so that an extension of the shaft of the tuning capacitor will line up with the dial. This places the box somewhat off center.

Power-supply components are mounted at the left-hand end of the chassis as viewed from the rear. The power transformer, plate and bias rectifiers, voltage-regulator tubes and filter choke L_5 are placed on the top side of the chassis. The

Fig. 6-84—The v.f.o. unit mounted in its cabinet. Holes are drilled in the dial cover to accommodate the switch shafts. At the right, a poker chip has been cemented to the v.f.o. set push-button switch so that it can be operated while tuning the v.f.o.; this makes frequency-spotting a one-handed operation.

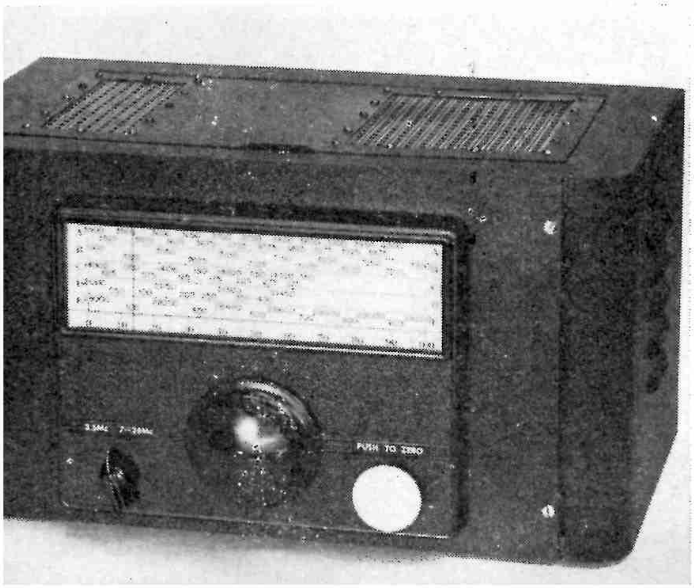
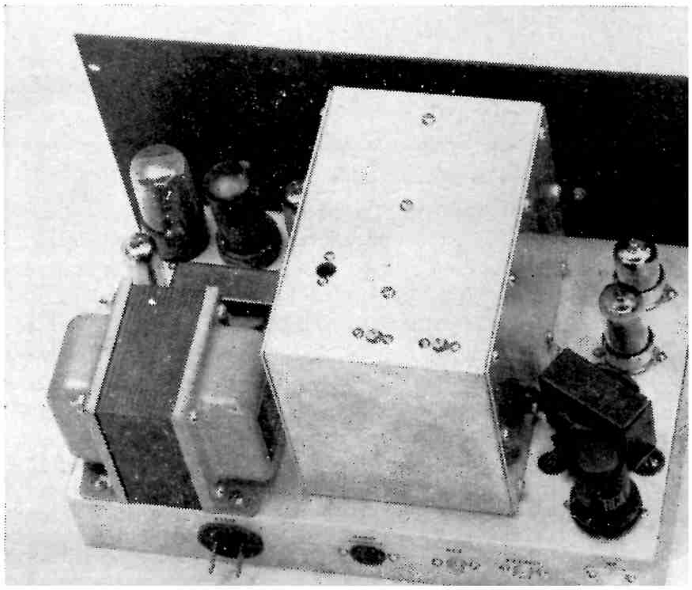


Fig. 6-85—Rear view of the v.f.o. unit. Power-supply components are to the left of the tuned-circuit compartment, and r.f. and 6J5 tubes to the right. The three screws along the center line of the box are used to fasten a stiffening strip of angle stock inside. Similar strips should be fastened against the side covers.



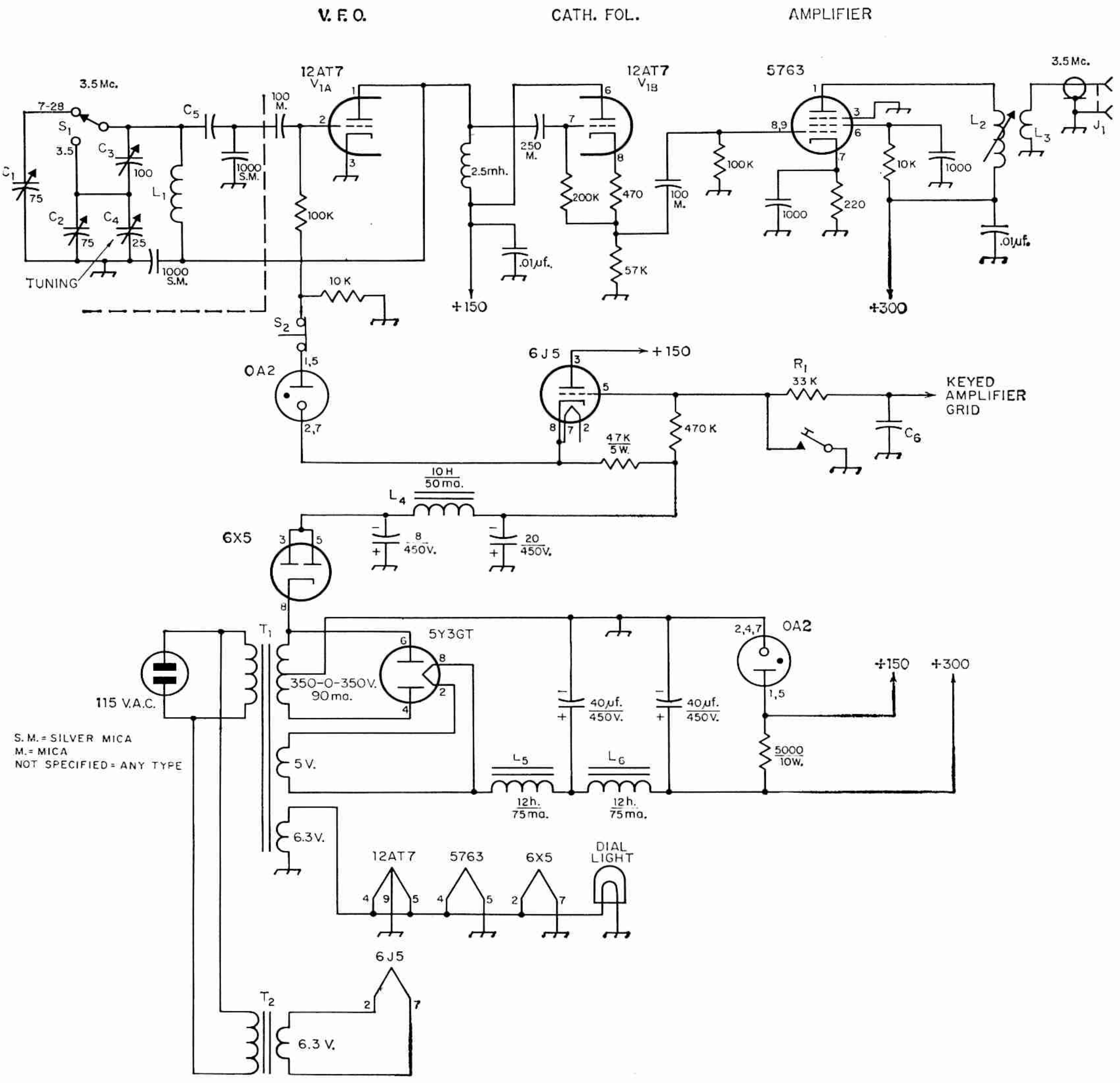


Fig. 6-86—Circuit diagram of the v.f.o., with its power supply and the keying system. Except as otherwise indicated, fixed resistors are ½ watt, capacitances are in μμf., resistances are in ohms. Capacitors marked with polarity are electrolytic.

C_1, C_2—75-μμf. variable (Hammarlund APC-75).

C_3—100-μμf. variable (Hammarlund APC-100).

C_4—25-μμf. variable (Millen 20025).

C_5—50-μμf. (Hammarlund APC-50); see text.

C_6—0.1-μf. 600-v. tubular, part of shaping circuit. Mounted in amplifier.

J_1—Coax connectors, chassis mounting.

L_1—30 turns No. 16, 1¾ inch diameter, 10 turns/inch (Airdux 1410T).

L_2—3.5 Mc.—72 turns No. 22 enam., close-wound on ⅜″ diameter slug-tuned form (Waters CSA-1012-1-WH).
7 Mc.—40 turns No. 22 close-wound on same form as above; 5-turn link.

L_3—10 turns, wound on cold end of, but insulated from, L_2.

L_4—10 hy., 50 ma. (Triad C-3X).

L_5, L_6—12 hy., 75 ma. (Triad C-5X).

R_1—33,000 ohms, part of shaping circuit. Mounted in amplifier.

S_1—Miniature rotary, 2-position (Centralab PA-2001).

S_2—Normally-closed push-button switch (Switchcraft 1002 modified with a longer shaft so as to extend through the main dial housing).

T_1—700 v. c.t., 90 ma.; 5 v., 3 amp.; 6.3 v., 3.5 amp. (Triad R-11A).

T_2—6.3-v. 0.6-ampere filament transformer.

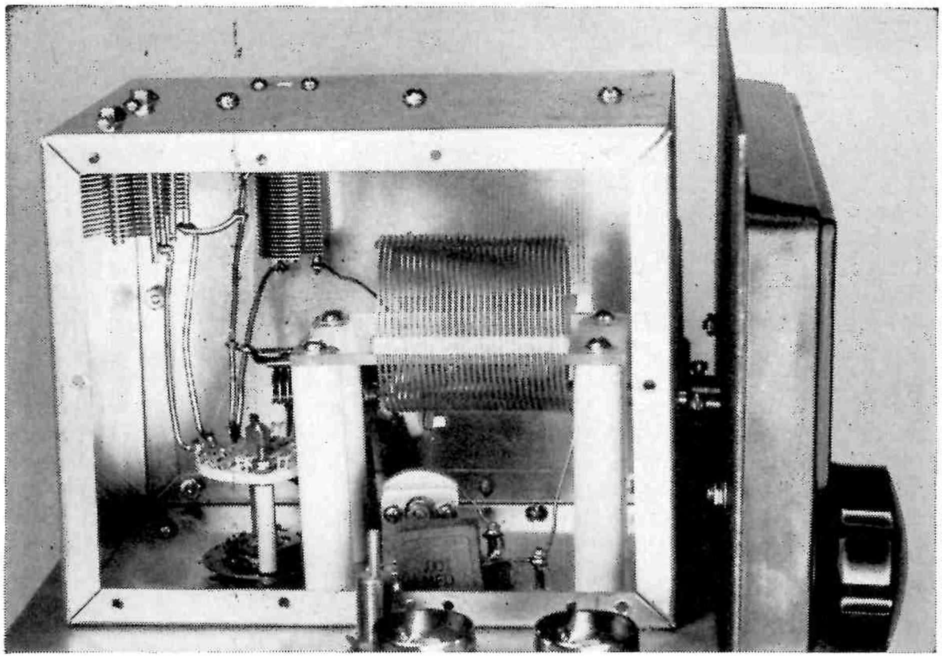


Fig. 6-87—The v.f.o. coil is mounted on ceramic pillars. The tuning capacitor C_4 can be seen behind the rear pair of insulators. The air capacitor C_5 is partially hidden by the 1000-$\mu\mu$f. silver mica capacitor below the coil. No. 14 wire is used between the switch and the coil and capacitors. In the foreground, transformer and tubes have been removed to show the adjusting screw of L_2.

bias filter choke, the plate filter choke L_6 and the filter capacitors are underneath. L_6 mounts with the same screws used for mounting L_5 above. Several ¼-inch holes should be drilled in the chassis in the vicinity of the power-supply components to help ventilate the under side of the chassis.

The v.f.o./cathode follower, amplifier and 6J5 tubes and their associated circuit components are at the left hand end of the chassis. The v.f.o. tube is close to the panel, followed by the 5763 amplifier, T_2 and 6J5 cathode follower. The slug-

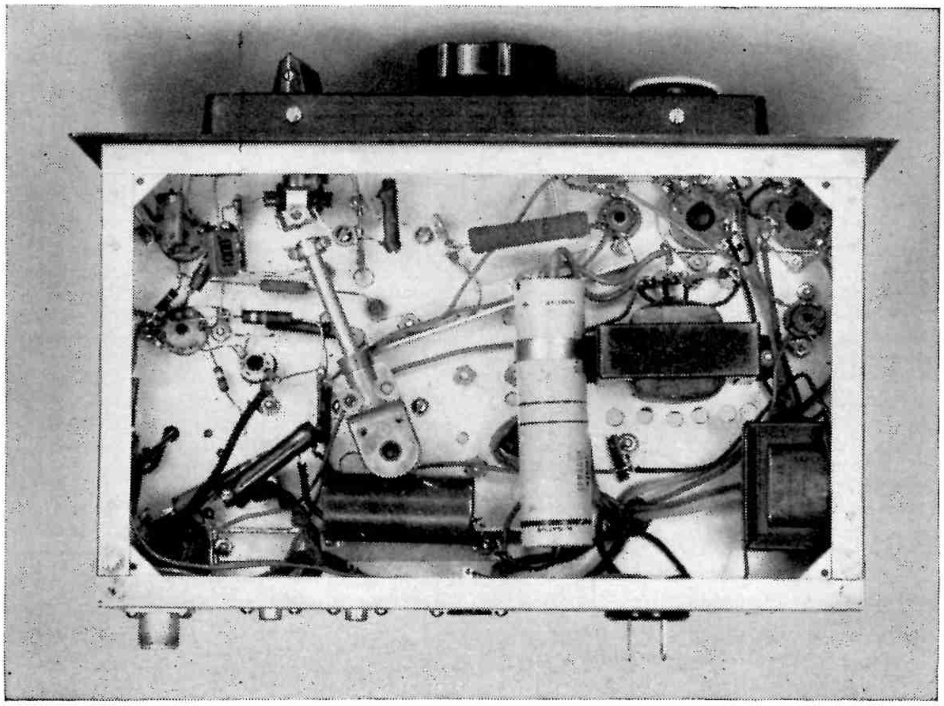


Fig. 6-88—Bottom view of the v.f.o. unit. The right-angle drive, right of center, drives the band-spread switch S_1. The small sections of aluminum angle stock are stiffeners added after the components were mounted. The method suggested in the text is preferable.

tuned coil L_2 is mounted alongside the 5763. It can be adjusted from the top of the chassis.

Along the rear edge of the chassis are a connector for the a.c. line, connectors for connecting a remote switch in parallel with S_2, for the key, for the keyed amplifier grid, and a coaxial connector for r.f. output.

Large rectangular ventilating holes are cut in the lid of the cabinet and then backed with patches of Reynolds perforated aluminum. If this detail is omitted, the temperature rise of the unit may cause considerable frequency drift.

Adjustment

In adjusting the v.f.o. frequency ranges, first set S_1 to the 80-meter position. With the dial set at zero (C_4 at maximum capacitance) adjust C_2 for a signal at 3500 kc. on a calibrated receiver. Then, with the dial of the v.f.o. set at the upper region of the scale, the signal should be heard at 4000 kc. If it is impossible to reach 4000 kc. with the v.f.o., the coil should be trimmed a part of a turn at a time.

In adjusting the second range (3500 to 3650 kc.), turn S_1 to the 7 — 28-Mc. position. Set C_3 temporarily at about half capacitance. Then, with the v.f.o. dial set at zero, adjust C_1 until a signal is heard at 3500 kc. Then check the v.f.o. frequency at the upper end of the dial. If the range does not go up to 3650 kc., C_3 should be increased a little and C_1 decreased to bring 3500 kc. at zero on the dial. If the tuning range goes above 3650 kc., C_3 should be decreased, and C_1 increased. A few trial settings should yield the correct range. The only other adjustment of the r.f. circuit is resonating the slug-tuned output coil. If set in the center of the tuning range, output should be reasonably constant over the entire range.

Adjustment of the keying circuit should be in accordance with the factors mentioned in Chapter Eight in connection with grid-block keying.

THE VACKAR VFO CIRCUIT

The Vackar variable-frequency oscillator appears to have some advantages over the usual Clapp circuit.[1] In the latter, the output amplitude varies greatly with frequency. In the Vackar circuit, the output varies only a little with frequency. The useful frequency range of the Clapp circuit is about 1.2 to 1; in the Vackar it is about 2.5 to 1. The first of these advantages should be of interest to amateurs.

My friend and colleague, Mr. James B. Ricks, W9TO, has pointed out that the 6AG7 is not the best tube to use for a series-tuned VFO; indeed the several papers originally describing these circuits invariably show triodes. The best tube is that one which has the lowest ratio of change of input capacitance to its mutual conductance. The operating mutual conductance for the cathode, control grid, and screen grid of a 6AG7 (as typically used as an oscillator) is low, despite its high value for the normal grid-to-plate circuitry. Also, it has a high input capacitance and high heater and plate power inputs. In consequence, this tube is not ideal for the purpose.

A small dual triode, the 12AT7, offers higher oscillator g_m in one triode section, lower input capacitance, and about one third the heater and plate power inputs required by the 6AG7. In consequence, it is a superior tube for series-tuned oscillators. The output voltage will be lower for the 12AT7, naturally, but a tube should not be evaluated for VFO use on the basis of power output.

W9TO has adapted the Vackar circuit to an amateur VFO with output on 80 meters using the 12AT7 in the circuit of Fig. 6-89. The first triode unit and its associated components form the oscillator proper; the other triode unit is a cathode follower which reduces loading effects on the oscillator frequency. Two of these VFO units have been made and tested; their frequency stability is excellent, and they key well. The output r.f. was measured as 1.2 volts r.m.s. using a General Radio v.t.v.m. The total current from the 255-volt regulated B supply was 16 ma., key down.

In series-tuned oscillators of the Clapp or Vackar type the characteristics of the series capacitor C_x are critical if the oscillator is to be keyed. An annoying chirp, slight but detectable, was finally traced to imperfection of this capacitor, even though it was a low temperature coefficient silvered mica one. Several silvered micas of good make were tried; they all produced slight chirp, some less than others. A so-called zero temperature coefficient (NPO) ceramic capacitor gave less chirp (very little, in fact), but the chirp was eliminated by using an APC air trimmer for C_x. Apparently, there is enough r.f. current through C_x to cause dielectric heating and a small resulting change in capacity even in these high-grade capacitors. This was confirmed indirectly by using for C_x a negative temperature coefficient (N750) ceramic capacitor. The chirp was tremendous!

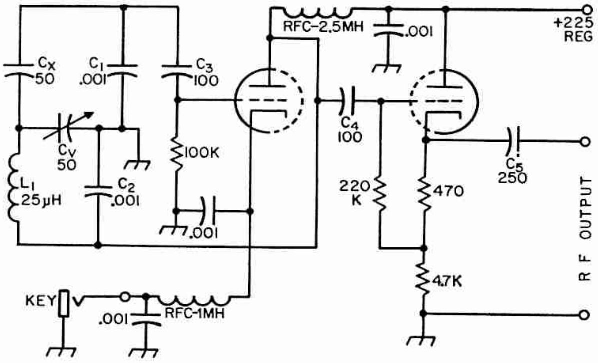


Fig. 6-89—Vackar series-tuned v.f.o. circuit at W9TO. The tube is a 12AT7 dual triode. R.f. output from the cathode-follower second section is 1.2 volts r.m.s.

C_1, C_2—Silver mica.
C_3, C_4, C_5—Mica
C —APC air variable.
Other copacitors are chramic.

Of course, the series capacitor is not the only possible cause of chirp; poor plate voltage regulation or a long time constant in the keying circuit might also contribute. To avoid this, the plate supply should be regulated, and series resistances and shunt capacitances in the keying circuit should be kept to a minimum.[2]

The circuit shown will key cleanly without chirp; with the constants shown it will be somewhat clicky, due to turning on and off rapidly; this makes it very desirable for use in a differential keying system in which the oscillator is turned on before the amplifier, and the amplifier is turned off before the oscillator. — *W9IK*

---

[1] Clapp, J. K., "Frequency Stable *LC* Oscillators," *Proc. of the I.R.E.*, Aug., 1954, Vol. 42, No. 8, page 1295.

[2] The chirp discussed in the preceding paragraph evidently is a slow one attributable to temperature effects. A chirp of the "dynamic" type often manifests itself as a click when the time constant of the keying circuit is very short, becoming observable as a chirp when key-thump elimination methods are used. — Ed.

This material originally appeared in *QST* for November, 1955. — ED.

Converting Surplus Transmitters for Novice Use

War-surplus radio equipment, available in many radio stores, is a good source of radio parts. Some of the transmitters and receivers can be made to operate in the amateur bands with little or no modification. It would be hard to find a more economical way for a Novice to get started on 40 or 80 meters than by adapting a normally-v.f.o.-controlled surplus "Command Set" to crystal control.

The "Command Sets" are parts of the SCR-274N and AN/ARC-5 equipments, transmitters and receivers designed for use in military aircraft. The two series are substantially identical in circuit and construction. Of the transmitters, two are of particular interest to the Novice. These are the BC-696 (part of 274N) or T19 (ARC-5) covering 3 to 4 Mc., and the BC-459 or T22, 7 to 9.1 Mc. The transmitter circuit consists of a 1626 triode variable-frequency oscillator that drives a pair of 1625s in parallel, which for Novice use can be run at 75 watts input. In addition to the 1626 and 1625s the transmitters include a 1629 magic-eye tube, which was used as a resonance indicator with a crystal for checking the dial calibration. The tubes have 12-volt heaters connected in series-parallel for 24-volt battery operation. The BC-696 and 459 are available from surplus dealers at prices ranging from five to fifteen dollars each, depending on condition.

Several methods have been described for converting the transmitters to crystal control for Novice use, but most of them didn't take into consideration the reconversion required to change back to v.f.o. when the Novice became a General-Class license holder.

In the modification to be described, the Novice requirement for crystal control is met by using a separate crystal-controlled oscillator. The output of the external oscillator is fed into the transmitter through a plug that fits into the 1626 oscillator socket. The 1626 is not used. The transmitter modifications are such that when it is desired to restore the transmitter to v.f.o. operation the external oscillator is unplugged and the 1626 is put back in its socket. No wiring changes are needed to go from crystal control to v.f.o.

In addition to the external oscillator, a power supply is required for the oscillator and transmitter (Fig. 6-90), and certain wiring changes are needed to make the transmitter itself suitable for amateur use. These changes consist primarily of removing two relays, changing the tube heater circuit for operation on 12 volts instead of 24 volts, and the addition of a power plug.

Transmitter Modifications

The 80- and 40-meter transmitters are practically identical except for frequency range, and the modifications are the same in both. Remove the top cover and bottom plate. Remove the tubes and crystal from their sockets so there will be no danger of breaking them as you work on the transmitter. If the sockets are not marked by tube types, mark them yourself so you'll know which tube goes where.

The following modifications are required:

1) Remove the antenna relay (front panel) and control relay (side of chassis) and unsolder and remove all wires that were connected to the relays with the exception of the wire going to Pin 4 on the oscillator socket.

2) Remove the wire-wound resistor mounted on the rear wall of the transmitter.

3) Unsolder the wire from Pin 7 of the 1629 socket and move it to Pin 2. Ground Pin 7.

4) Unsolder the wires from Pin 1 of the 1625 closest to the drive shaft for the variable capacitors and solder the wires to Pin 7. Run a lead from the same Pin 1 to the nearest chassis ground.

5) Unsolder all leads from the power socket at the rear of the chassis and remove the socket. The socket can be pried off with a screwdriver.

6) Unsolder the end of the 20-ohm resistor (red-black-black) that is connected to Pin 4 on the oscillator socket and connect it to Pin 6

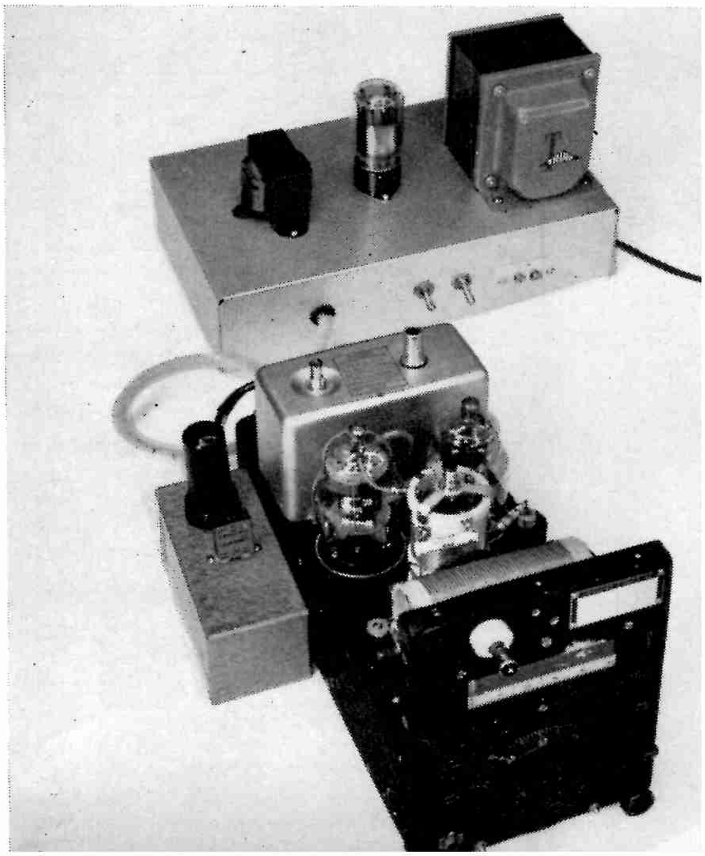


Fig. 6-90—The complete Novice setup, in this case using the 80-meter (BC-457) transmitter. Note the key jack at the lower-left corner of the transmitter panel. The crystal oscillator is connected to the transmitter oscillator-tube socket with a short length of cable terminating in an octal plug. A small notch should be cut in the transmitter cover to provide clearance for the cable when the cover is installed. The power transformer, rectifier, and choke are mounted on top of the power-supply chassis at the rear, and the control switches are mounted on the wall as shown. Remaining components are underneath.

of the calibration crystal socket. There is also a lead on Pin 4 that was connected to the keying relay; connect this lead to the nearest chassis ground point.

7) Mount an octal socket (Amphenol 78-RS8) in the hole formerly occupied by the power socket. Install a solder lug under one of the nuts holding the socket mounting.

8) Wire the octal socket as shown in Fig. 6-91. One of the leads unsoldered from the original power socket is red with a white tracer. This is the B+ lead for the 1625s. The yellow lead is the screen lead for the 1625s and the white lead is the heater lead. Although the manuals covering this equipment specify these colors, it's safer not to take them for granted; check where each lead actually goes before connecting it to the new power socket. The lead from Pin 1 on the power socket to Pin 6 on the calibration-crystal socket is the oscillator plate-voltage lead. The leads from Pins 7 and 8 on the power plug to Pins 1 and 6 on the oscillator socket are new leads to carry power to the external crystal-controlled oscillator. The lead from Pin 4 of the power socket to Pin 2 on the 1629 (resonance indicator) socket is the 12-volt heater lead.

9) Mount a closed-circuit phone jack at the lower left-hand corner of the front panel. Connect a lead from the ungrounded phone jack terminal to Pin 6 (cathode) of either of the 1625 sockets. This completes the modification.

Crystal-Controlled Oscillator Details

The external crystal-controlled oscillator circuit, shown in Fig. 6-92, uses a 6AG7 in the grid-plate oscillator circuit. Either 80- or 40-meter crystals are required, depending on the band in use, A tuned plate circuit is not required in the oscillator; it was found that more than adequate grid drive could be obtained with the setup as shown.

Output from the oscillator is fed to the transmitter through an 8-inch length of RG-58 coax cable. The cable is terminated in an octal plug, P_2, which is plugged into the oscillator tube socket in the transmitter. Power for the external oscillator is obtained through this socket.

The crystal-controlled oscillator is built in and on a 4 × 2 × 2¾-inch aluminum box. The tube and crystal sockets are mounted on top of the box and the remaining components inside. Layout of parts is not particularly critical but the general arrangement shown in Figs. 6-90 and 6-93 should be followed to insure good results.

In the completed setup, oscillator and amplifier, the cathodes of the 1625s are keyed and the crystal oscillator runs continuously during transmissions. It is thus necessary to turn the oscillator off during standby periods, and this is accomplished by opening the B-plus switch on the power supply. This method is used in preference to keying the oscillator and amplifier simultaneously because keying the oscillator is likely to make the signal chirpy. With amplifier keying the signal is a real T9X.

Power Supply

Fig. 6-91 shows the circuit of the power supply, which uses a 5U4G rectifier and a capacitor-input filter. The power transformer, T_1, is a type made by several manufacturers. To obtain the necessary 12.6 volts for the heaters, a 6.3-volt filament transformer is connected in series with the 6.3-volt winding on T_1. This setup also will provide 6.3 volts for the heater of the 6AG7. Current requirement for the 6AG7 heater is 0.65 amp. and for the 1625s, 0.9 amp. total.

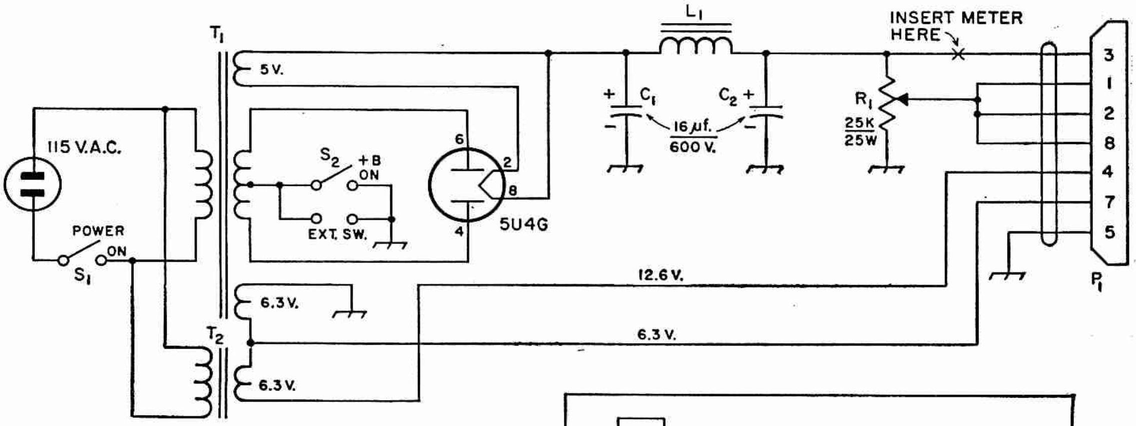


Fig. 6-91—Circuit diagram of power socket and power supply.

C_1, C_2—16-μf., 600-volt electrolytic (Sprague TVA-1965, Aerovox PRS).

J_1—Octal socket (Amphenol 78-RS8).

L_1—1- to 2-hy., 200-ma. filter choke, TV replacement type (Stancor C2325 or C2327, or equivalent).

P_1—Octal cable plug (Amphenol 86-PM8).

R_1—25,000 ohms, 25 watts, with slider.

S_1, S_2—Single-pole, single-throw toggle switch.

T_1—Power transformer, 100 volts center-tapped, 200 ma.; 5 volts, 3 amp.; 6.3 volts, 6 amp. (Knight 61G414, Triad R-21A, or equivalent).

J_1	1	PIN 6, CRYSTAL SOCKET (OSC. B+).
	2	PIN 5 1625 SOCKET (SCREEN).
	3	PLATE LEAD OF 1625s.
	4	PIN 2 1629 SOCKET (12.6 V.A.C.)
	5	CHASSIS GROUND.
	6	NO CONNECTION.
	7	PIN 6 1626 SOCKET (6.3 V.A.C.).
	8	PIN 1 1626 SOCKET (XTAL OSC. B+).

CONNECTIONS TO
POWER SOCKET ON TRANSMITTER.

T_2—Filament transformer, 6.3 volts, 3 amp. (Triad F-16X, Knight 62-G-031, or equivalent).

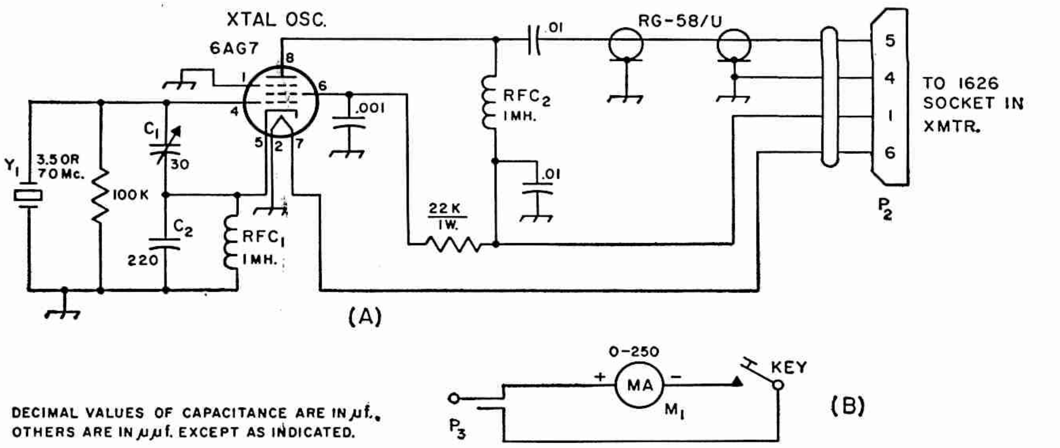


Fig. 6-92—(A) Circuit diagram of external crystal-controlled oscillator. Unless otherwise specified, resistances are in ohms, resistors are ½ watt. The 0.01- and 0.001-μf. capacitors are disk ceramic. (B) Method of connecting the milliammeter in series with the key.

C_1—3-30-μμf. trimmer.
C_2—220-μμf. fixed mica.
M_1—0-250 d.c. milliammeter.
P_2—Octal plug, male (Amphenol 86-PM8).
P_3—Phone plug.
RFC_1, RFC_2—1-mh. r.f. chokes.
Y_1—3.5- or 7-Mc. Novice-band crystal, as required.

To turn off the plate voltages on the transmitter during stand-by periods, the center tap of T_1 is opened. This can be done in two ways; by S_2, or by a remotely-mounted switch whose leads are connected in parallel with S_2. A two-terminal strip is mounted on the power-supply chassis, the terminals being connected to S_2 which is also on the chassis. The remotely-mounted switch can be installed in any convenient location at the operating position. A single-pole, single-throw switch can be used for this purpose or, if desired, a multicontact switch can be used to perform simultaneously this and other functions, such as controlling an antenna-changeover relay.

The high-voltage and heater leads are brought out in a cable to an octal plug, P_1, that connects to J_1 on the transmitter. The length of the cable will, of course, depend on where you want to install the power supply. Some amateurs prefer to have the supply on the floor under the operating desk rather than have it take up room at the operating position.

The supply shown here was constructed on a 3 × 6 × 10-inch chassis. The layout is not critical, nor are there any special precautions to take during construction other than to observe polarity in wiring the electrolytic capacitors and to see that the power leads are properly insulated. Never have P_1 unplugged from J_1 when the power supply is turned on; there is danger of electrical shock at several pins of P_1. Interchanging the inserts of P_1 and J_1 will remove this hazard.

When wiring P_1 don't connect the B-plus lines to Pins 2 or 3, the amplifier plates and screens, at first. It is more convenient to test the oscillator without plate and screen voltages on the amplifier.

When the supply is completed, check between chassis ground and the 12.6-volt lead with an a.c. voltmeter to see if the two 6.3-volt windings are connected correctly. If you find that the voltage is zero, reverse one of the windings. If you don't have an a.c. meter you can check by observing the heaters in the 1625s. They will light up if you have the windings connected correctly. Incidentally, leave B plus off, by opening S_2, for this check.

Next, set the slider on the bleeder resistor, R_1, at about one-quarter of the total resistor length, measured from the B-plus end of the bleeder. Be sure to turn off the power when making this adjustment. With the tap set about one-quarter of the way from the B-plus end of the bleeder the oscillator plate and amplifier screen voltages will be approximately 250 volts.

Testing the Transmitter

A key and meter connected as shown in Fig. 6-92 are needed for checking the transmitter. When P_3 is plugged into the jack in the transmitter it will measure the cathode current of the 1625s. The cathode current is the sum of the plate, screen and control-grid currents. Some amateurs prefer to install the meter in the plate lead so it reads plate current only. This can be done by opening the B-plus line at the point marked "X" in Fig. 6-91, and inserting the meter in series with the line. However, unless more than one meter is available, don't install it in the power supply setup in this way until after the tests described below have been made.

Insert the external oscillator plug, P_2, into the 1626 socket and connect P_1 to the transmitter. Plug P_1 into the key jack on the front panel of the transmitter. With S_2 open, turn on the power and allow a minute or two for the tubes to warm up. Next, close the center-tap connection, S_2, on the power transformer. Set the transmitter dial to the same frequency as that of the crystal in use and close the key. A slight indication of grid current should show on the meter. There is no plate or screen current because there are no screen or plate voltages on the amplifier. If no grid current is obtained, adjust C_1 to the point where grid current shows, or try another crystal.

The next step is to peak the amplifier grid circuit — that is, the 1626 v.f.o. tank — for maximum grid-current reading. The v.f.o. trimmer capacitor is in an aluminum box on the top of the chassis at the rear. There is a ½-inch diameter hole in the side of the box; loosen the small screw visible through this hole, thus unlocking the rotor shaft of the trimmer capacitor. Move the rotor-arm shaft in either direction, observing the meter reading, and find the position that gives the highest reading. This should be something more than 10 ma.

Now connect the plate and screen voltage leads to P_1. Be sure to turn off the power supply before making the connections!

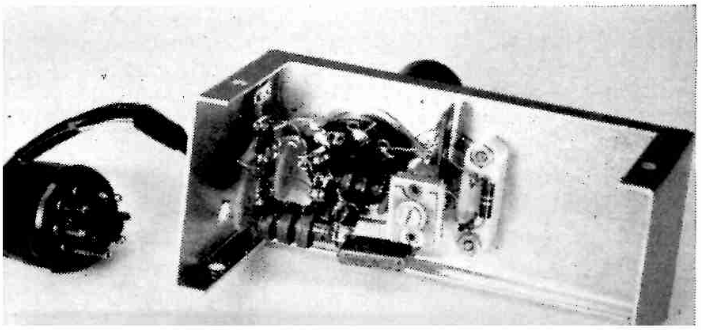
The first test of the rig should be with a dummy load; a 115-volt, 60-watt light bulb can be used for this purpose. The lamp should be connected between the antenna terminal and chassis ground. However, to make the lamp take power it may be necessary to add capacitance in parallel with it. A receiving-type variable capacitor having 250 $\mu\mu$f. or more maximum capacitance will be adequate for the job.

Turn on the power and allow the tubes to warm up, but leave the key open. Set the antenna coupling control on the transmitter to 7 or 8, and set the variable capacitor connected across the dummy load to about maximum capacitance. Next, close the key and adjust the antenna inductance control for an increase in cathode current. Turn the frequency control for a dip in current reading. The indicated frequency will probably differ from that of the crystal in use, but don't worry about it.

Adjust the three transmitter controls, antenna inductance, antenna coupling, and frequency, along with the variable capacitor across the lamp load, until the lamp lights up to apparently full brilliance. The cathode current should be between 150 and 200 ma. With the transmitter fully loaded, adjust C_1 in the crystal oscillator so that the lamp brilliance just starts to decrease. This is the optimum setting for C_1 and it can be left at this setting, no further adjustments being required.

If a d.c. voltmeter is available, check the different voltages in the setup. Using the power supply shown here, the plate voltage on the 1625s is approximately 400 with the amplifier fully loaded. With the plate voltage on the oscillator and screen voltage on the 1625s adjusted to 250 volts (tap on R_1), the oscillator screen voltage is 160 volts. The oscillator takes approximately 30 ma. and the 1625 amplifier screens about 10 ma. when the amplifier is fully loaded.

Fig. 6-93—This bottom view of the crystal oscillator shows the arrangement of components. Terminal strips are used for the cable connections and also as a support for C_1, the feedback capacitor.



Getting on the Air

To put the transmitter on the air it is necessary only to connect an antenna to the antenna post and connect a ground lead from the transmitter chassis to a water-pipe ground or to a metal stake driven in the ground. Almost any length of antenna will work, but for best results the minimum length should not be less than about ⅛ wavelength for the band in use. This is approximately 33 feet for 80 meters and 16 feet for 40 meters. It is of course better to make the antenna longer — and to be sure to get the far end as high as possible.

An output indicator will prove to be a handy device for knowing when power is actually going into the antenna. For this purpose use a 6.3-volt, 150-ma. dial lamp. Connect two leads, each about one foot long, to the shell and base of the bulb, respectively. Clip one lead to the antenna post and the other lead on the antenna wire two feet from antenna post. A small amount of power will go through the bulb and this will provide a visual indication of output. Follow the same tuning procedure as outlined above for the dummy antenna. If the bulb gets so bright that it is in danger of burning out, move the leads closer together to reduce the pickup.

It may be found that certain antenna lengths won't work — that is, the amplifier won't load — no matter where the antenna coupling and inductance are set. In such a case, connect a variable capacitor — like the one used with the lamp dummy — between the antenna post and the transmitter chassis. Adjust the capacitor and antenna inductance for maximum brilliance of the output indicator; this will be the best setting for the controls.

A superior antenna system uses a two-wire feeder system and an antenna coupler; examples are given in Chapters 13 and 14. If a coupler is used, the transmitter and coupler should be connected together with coax line. The inner conductor of the coax should be connected to the antenna terminal and the outer braid to the transmitter case, as close to the antenna terminal as possible. If desired, the antenna terminal can be removed and a coax fitting substituted.

When the coveted General Class ticket is obtained, it is only necessary to unplug the crystal oscillator, put the original tube back in the rig, and move out of the Novice band.

CHAPTER 7

Power Supplies

Essentially pure direct-current plate supply is required to prevent serious hum in the output of receivers, speech amplifiers, modulators and transmitters. In the case of transmitters, pure d.c. plate supply is also dictated by government regulation.

The filaments of tubes in a transmitter or modulator usually may be operated from a.c. However, the filament power for tubes in a receiver (excepting power audio tubes), or those in a speech amplifier may be a.c. only if the tubes are of the indirectly-heated-cathode type, if hum is to be avoided.

Wherever commercial a.c. lines are available, high-voltage d.c. plate supply is most cheaply and conveniently obtained by the use of a transformer-rectifier-filter system. An example of such a system is shown in Fig. 7–1.

In this circuit, the plate transformer, T_1, steps up the a.c. line voltage to the required high voltage. The a.c. is changed to pulsating d.c. by the rectifiers, V_1 and V_2. Pulsations in the d.c. appearing at the output of the rectifier (points A and B) are smoothed out by the filter composed of L_1 and C_1. R_1 is a *bleeder* resistor. Its chief function is to discharge C_1, as a safety measure, after the supply is turned off. By proper selection of value, R_1 also helps to minimize changes in output voltage with changes in the amount of current drawn from the supply. T_2 is a step-down transformer to provide filament voltage for the rectifier tubes. It must have sufficient insulation between the filament winding and the core and primary winding to withstand the peak value of the rectified voltage. T_3 is a similar transformer to supply the filaments or heaters of the tubes in the equipment operating from the supply. Frequently, these three transformers are combined in a single unit having a single 115-volt primary winding and the required three secondary windings on one core.

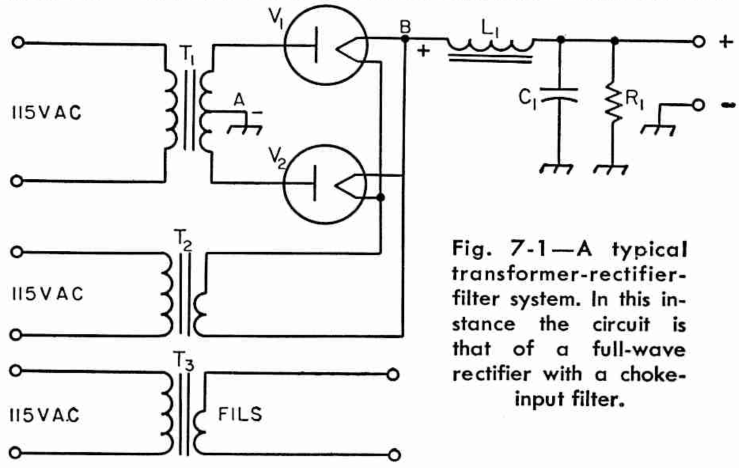


Fig. 7-1—A typical transformer-rectifier-filter system. In this instance the circuit is that of a full-wave rectifier with a choke-input filter.

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 (or filament), 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 in (A) 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 (one pulsation per cycle), 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), for the same d.c. power output, 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 is required with the circuit. When the plate of V_1 is positive, current flows through the load to the center tap. Current cannot flow through V_2 because at this instant its cathode (or filament) is positive in respect to its plate. When the polarity reverses, V_2 conducts and current again flows through the load to the center-tap, this time through V_2.

The average output voltage is 0.45 times the r.m.s. voltage of the entire transformer-secondary, or 0.9 times the voltage across *half* of the transformer secondary. For the same *total* secondary voltage, the average output voltage is the same as that delivered with a half-wave rectifier. However, as can be seen from the sketches of the output wave form in (B) to the right, 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 rating of each rectifier need be only half the total load current drawn from the supply.

Two separate transformers, with their primaries connected in parallel and secondaries connected in series (with the proper polarity) may be used in this circuit. However, if this substitution is made, the primary volt-ampere rating must be reduced to about 40 per cent less than twice the rating of one transformer.

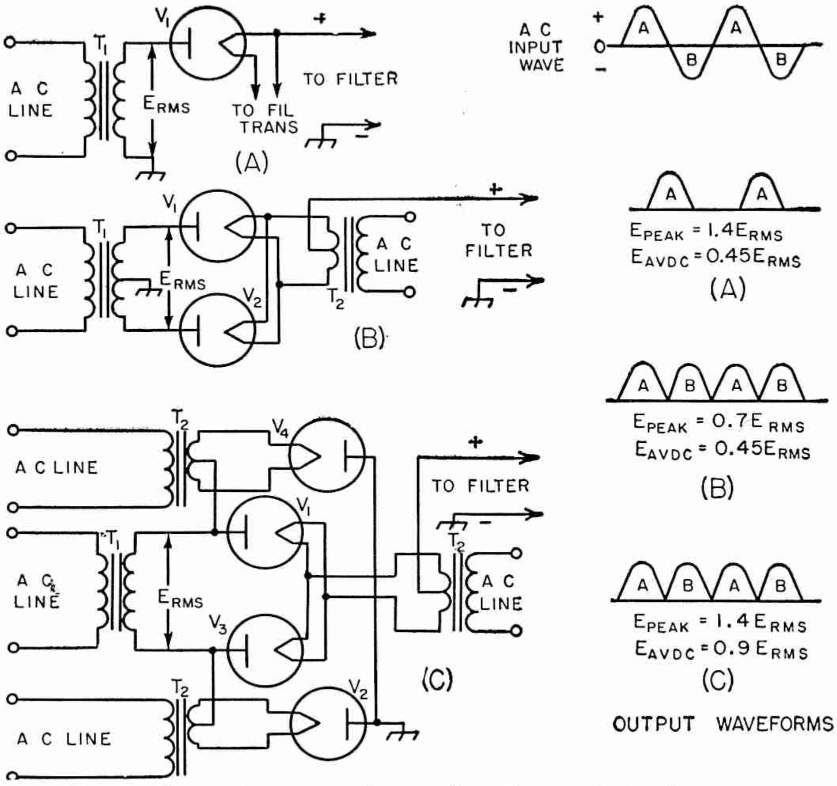


Fig. 7-2—Fundamental vacuum-tube rectifier circuits. A—Half-wave. B—Full-wave. C—Full-wave bridge. A.c.-input and pulsating-d.c. output wave forms are shown at the right. Output-voltage values indicated do not include rectifier drops. Other types of rectifiers may be substituted.

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 V_1, through the load and thence through V_2. During this period current cannot flow through rectifier V_4 because its plate is negative with respect to its cathode (or filament). Over the other half of the cycle, current flows through V_3, through the load and thence through V_4. Three filament transformers are needed — one for V_1 and V_3 and one each for V_2 and V_4. The output wave shape (C), to the right, 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. Each rectifier in a bridge circuit should have a minimum load-current rating of one half the total load current to be drawn from the supply.

Rectifiers

High-Vacuum Rectifiers

High-vacuum rectifiers depend entirely upon the thermionic emission from a heated filament 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 275 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, two tubes being required for a full-wave rectifier circuit. A few of the lower-voltage types have indirectly heated cathodes, but are limited in heater-to-cathode voltage rating.

Mercury-Vapor Rectifiers

The voltage drop through a mercury-vapor rectifier is practically constant at approximately 15 volts regardless of the load current. 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 nearby receivers, sometimes difficult to eliminate. R.f. filtering in the primary circuit and at the rectifier plates as well as shielding may be required. 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 and Other Semiconductor Rectifiers

Selenium, germanium and silicon rectifiers are finding increasing application in power supplies for amateur equipment. These units have the advantages of compactness, low internal voltage drop (about 5 volts per unit) and low operating temperature. Also, no filament transformers are required.

Individual units of all three types are available with input ratings of 130 volts r.m.s. Selenium units are rated at up to 1000 ma. or more d.c. load current; germanium units have ratings up to 400 ma., and silicon units up to 500 ma. In full-wave circuits these load-current figures can be doubled.

The extreme compactness of silicon types makes feasible the stacking of several units in series for higher voltages. Standard stacks are available that will handle up to 2000 volts r.m.s. input at a d.c. load current of 325 ma. Two of these stacks in a full-wave circuit will handle 650 ma., although they are comparatively expensive.

Semiconductor rectifiers may be substituted in any of the basic circuits shown in Fig. 7-2, the terminal marked "+" or "cathode" corresponding to the filament connection. Advantage may be taken of the voltage-multiplying circuits discussed in a later section of this chapter in adapting rectifiers of this type.

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. This is sometimes dependent on whether a choke- or capacitive-input filter is used. Others, particularly mercury-vapor types, are rated according to maximum inverse peak voltage — the peak voltage between plate and cathode while the tube is not conducting. In the circuits of 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, except that if a capacitive-input filter is used with the half-wave rectifier circuit of Fig. 7-2A, the multiplying factor becomes 2.8.

All rectifier tubes are rated also as to maximum d.c. load current and many, in addition, carry peak-current ratings, all of which should be carefully observed to assure normal tube life. With a capacitive-input filter, the peak current may run several times the d.c. current, while with a choke-input filter the peak value may not run more than twice the d.c. load 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 space surrounding them to provide for ventilation. When mercury-vapor tubes are first placed in service, and each time after the mercury has been disturbed, as by removal from the socket to a horizontal position, they should be run with filament voltage only for 30 minutes before applying high voltage. After that, a delay of 30 seconds is recommended each time the filament is turned on.

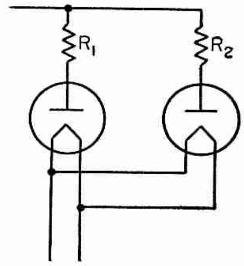


Fig. 7-3—Connecting mercury-vapor rectifiers in parallel for heavier currents. R_1 and R_2 should have the same value, between 50 and 100 ohms, and corresponding filament terminals should be connected together.

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. With mercury-vapor types, equalizing resistors of 50 to 100 ohms should be connected in series with each plate, as shown in Fig. 7-3, to help maintain an equal division of current between the two rectifiers.

Filters

The pulsating d.c. waves from the rectifiers shown in Fig. 7-2 are not sufficiently constant in amplitude to prevent hum corresponding to the pulsations. Filters consisting of capacitances and inductances are required between the rectifier and the load to smooth out the pulsations to an essentially constant d.c. voltage. Also, upon the design of the filter depends to a large extent the d.c. voltage output, the *voltage regulation* of the power supply and the maximum load current that can be drawn from the supply without exceeding the peak-current rating of the rectifier.

Power-supply filters fall into two classifications, depending upon whether the first filter element following the rectifier is a capacitor or a choke. Capacitive-input filters are characterized by relatively high output voltage in respect to the transformer voltage, but poor voltage regulation. Choke-input filters result in much better regulation, when properly designed, but the output voltage is less than would be obtained with a capacitive-input filter from the same transformer.

Voltage Regulation

The output voltage of a power supply always decreases as more current is drawn, not only because of increased voltage drops in the transformer, filter chokes and the rectifier (if high-vacuum rectifiers are used) but also because the output voltage at light loads tends to soar to the peak value of the transformer voltage as a result of charging the first capacitor. By proper filter design the latter effect can be eliminated. The change in output voltage with load is called *voltage regulation* and is expressed as a percentage.

$$\textit{Per cent regulation} = \frac{100\ (E_1 - E_2)}{E_2}$$

Example: No-load voltage = E_1 = 1550 volts.
Full-load voltage = E_2 = 1230 volts.

$$\text{Percentage regulation} = \frac{100\ (1550 - 1230)}{1230}$$

$$= \frac{32{,}000}{1230} = 26 \text{ per cent.}$$

Regulation may be as great as 100% or more with a capacitive-input filter, but by proper design can be held to 20% or less with a choke-input filter.

Good regulation is desirable if the load current varies during operation, as in a keyed stage or a Class B modulator, because a large change in voltage may increase the tendency toward key clicks in the former case or distortion in the latter. On the other hand, a steady load, such as is represented by a receiver, speech amplifier or unkeyed stages in a transmitter, does not require good regulation so long as the proper voltage is obtained under load conditions. Another consideration that makes good voltage regulation desirable is that the filter capacitors must have a voltage rating safe for the highest value to which the voltage will soar when the external load is removed.

When essentially constant voltage, regardless of current variation is required (for stabilizing an oscillator, for example), special voltage-regulating circuits described elsewhere in this chapter are used.

Load Resistance

In discussing the performance of power-supply filters, it is sometimes convenient to express the load connected to the output terminals of the supply in terms of resistance. The load resistance is equal to the output voltage divided by the total current drawn, including the current drawn by the bleeder resistor.

Input Resistance

The sum of the transformer impedance and the rectifier resistance is called the input resistance. The approximate transformer impedance is given by

$$Z_{\mathrm{TR}} = N^2 R_{\mathrm{PRI}} + R_{\mathrm{SEC}}$$

where N is the transformer turns ratio, primary to secondary (primary to ½ secondary in the case of a full-wave rectifier), and R_{PRI} and R_{SEC} are the primary and secondary resistances respectively. R_{SEC} will be the resistance of half of the secondary in the case of a full-wave circuit.

Bleeder

A bleeder resistor is a resistance connected across the output terminals of the power supply (see Fig. 7-1). Its functions are to discharge the filter capacitors as a safety measure when the power is turned off and to improve voltage regulation by providing a minimum load resistance. When voltage regulation is not of importance, the resistance may be as high as 100 ohms per volt. The resistance value to be used for voltage-regulating purposes is discussed in later sections. From the consideration of safety, the power rating of the resistor should be as conservative as possible, since a burned-out bleeder resistor is more dangerous than none at all!

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 capacitors 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, the output ripple from the power supply should not exceed 5 per cent. The ripple in the output of supplies for voice transmitters should not exceed 1 per cent. Class B modulators require a ripple reduction to about 0.25%, while v.f.o.'s, high-gain speech amplifiers, and receivers may require a reduction in ripple to 0.01%.

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.

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 lowered.

● CAPACITIVE-INPUT FILTERS

Capacitive-input filter systems are shown in Fig. 7-4. Disregarding voltage drops in the chokes, all have the same characteristics except in respect to ripple. Better ripple reduction will be obtained when *LC* sections are added, as shown in Figs. 7-4B and C.

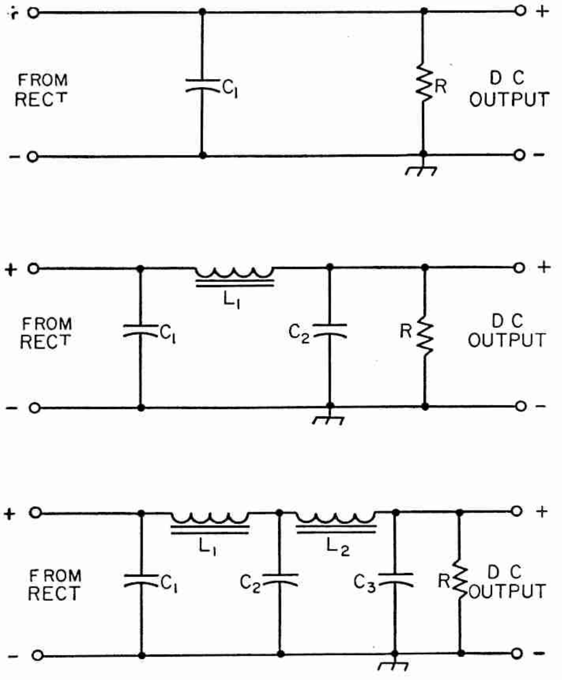


Fig. 7-4—Capacitive-input filter circuits. A—Simple capacative. B—Single-section. C—Double-section.

Output Voltage

To determine the approximate d.c. voltage output when a capacitive-input filter is used, reference should be made to the graph of Fig. 7-5.

Example:
Transformer r.m.s. voltage — 350
Input resistance — 200 ohms
Maximum load current, including bleeder current — 175 ma.

$$\text{Load resistance} = \frac{350}{0.175} = 2000 \text{ ohms approx.}$$

From Fig. 7-5, for a load resistance of 2000 ohms and an input resistance of 200 ohms, the d.c. output voltage is given as slightly over 1 times the transformer r.m.s. voltage, or about 350 volts.

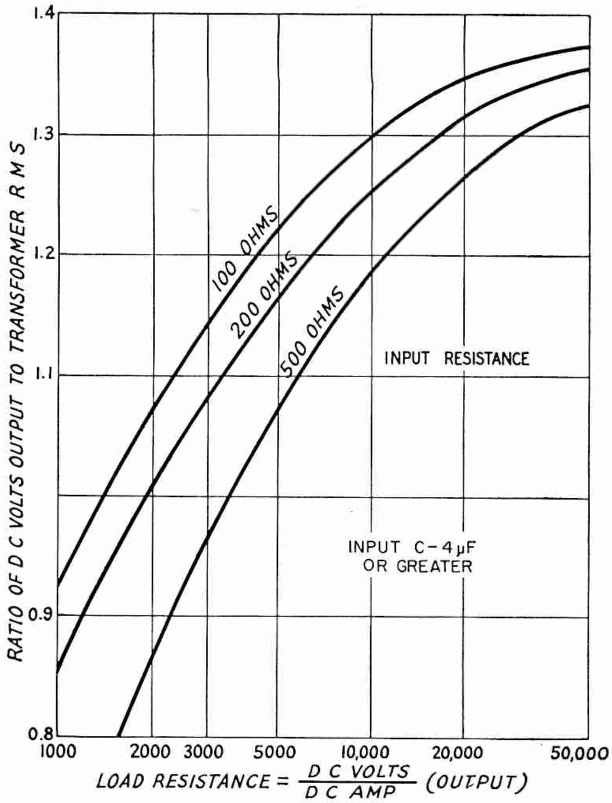


Fig. 7-5—Chart showing approximate ratio of d.c. output voltage across filter input capacitor to transformer r.m.s. secondary voltage for different load and input resistances.

Regulation

If a bleeder resistance of 50,000 ohms is used, the d.c. output voltage, as shown in Fig. 7-5, will rise to about 1.35 times the transformer r.m.s. value, or about 470 volts, when the external load is removed. For greater accuracy, the voltage drops through the input resistance and the resistance of the chokes should be subtracted from the values determined above. For best regulation with a capacitive-input filter, the bleeder resistance should be as low as possible without exceeding the transformer, rectifier or choke ratings when the external load is connected.

Maximum Rectifier Current

The maximum current that can be drawn from a supply with a capacitive-input filter without exceeding the peak-current rating of the rectifier may be estimated from the graph of Fig. 7-6. Using values from the preceding example, the ratio of peak rectifier current to d.c. load current for 2000 ohms, as shown in Fig. 7-6 is 3. Therefore, the maximum load current that can be drawn without exceeding the rectifier rating is ⅓ the peak rating of the rectifier. For a load current of 175 ma., as above, the rectifier peak current rating should be at least 3 × 175 = 525 ma.

With bleeder current only, Fig. 7-6 shows that the ratio will increase to over 8. But since the bleeder draws less than 10 ma. d.c., the rectifier peak current will be only 90 ma. or less.

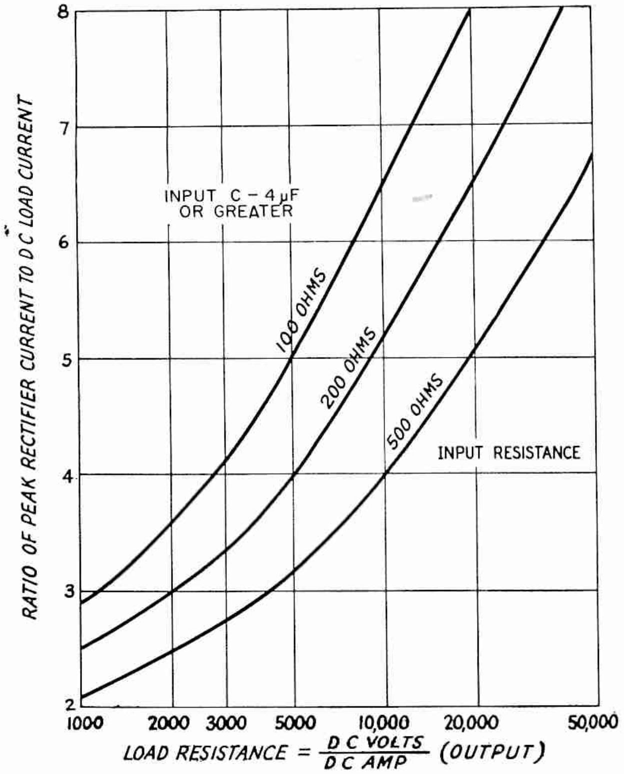


Fig. 7-6—Graph showing the relationship between the d.c. load current and the rectifier peak plate current with capacitive input for various values of load and input resistance.

Ripple Filtering

The approximate ripple percentage after the simple capacitive filter of Fig. 7-4A may be determined from Fig. 7-7. With a load resistance of 2000 ohms, for instance, the ripple will be approximately 10% with an 8-µf. capacitor or 20% with a 4-µf. capacitor. For other capacitances, the ripple will be in inverse proportion to the capacitance, e.g., 5% with 16 µf., 40% with 2 µf., and so forth.

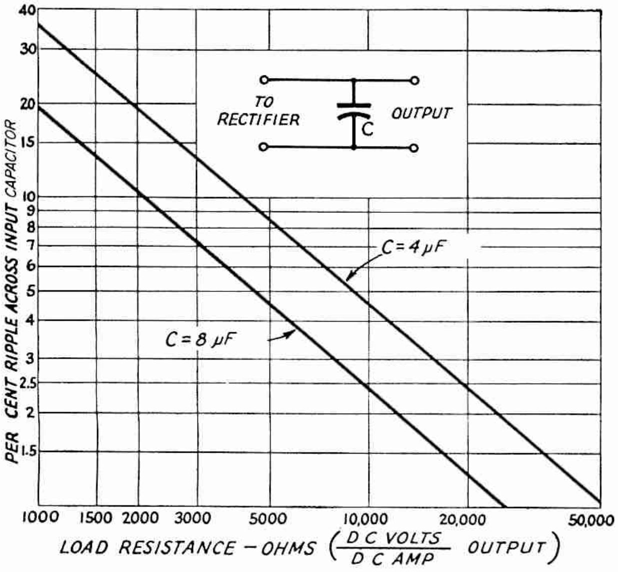


Fig. 7-7—Showing approximate 120-cycle percentage ripple across filter input capacitor for various loads.

The ripple can be reduced further by the addition of *LC* sections as shown in Figs. 7-4B and C. Fig. 7-8 shows the factor by which the ripple from any preceding section is reduced depending on the product of the capacitance and inductance added. For instance, if a section composed of a choke of 5 h. and a capacitor of 4 µf. were to be added to the simple capacitor of Fig. 7-4A, the product is 4 × 5 = 20. Fig. 7-8 shows that the original ripple (10% as above with 8 µf. for example) will be reduced by a factor of about 0.08. Therefore the ripple percentage after the new section will be approximately 0.08 × 10 = 0.8%. If another section is added to the filter, its reduction factor from Fig. 7-8 will be applied to the 0.8% from the preceding section; 0.8 × 0.08 = 0.064% (if the second section has the same *LC* product as the first).

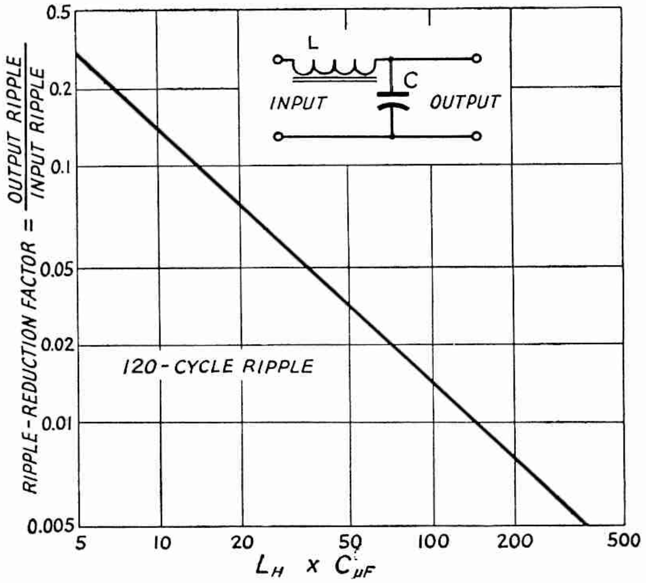


Fig. 7-8—Ripple-reduction factor for various values of *L* and C in filter section. *Output ripple = input ripple × ripple factor.*

CHOKE-INPUT FILTERS

Much better voltage regulation results when a choke-input filter, as shown in Fig. 7-9, is used. Choke input also permits better utilization of the rectifier, since a higher load current usually can be drawn without exceeding the peak current rating of the rectifier.

Minimum Choke Inductance

A choke-input filter will tend to act as a capacitive-input filter unless the input choke has at least a certain minimum value of inductance called the **critical** value. This critical value is given by

$$L_{\mathrm{h}} = \frac{E_{\mathrm{VOLTS}}}{I_{\mathrm{MA.}}}$$

where *E* is the output voltage of the supply, and *I* is the current being drawn from the supply.

If the choke has at least the critical value, the output voltage will be limited to the average value of the rectified wave at the input to the

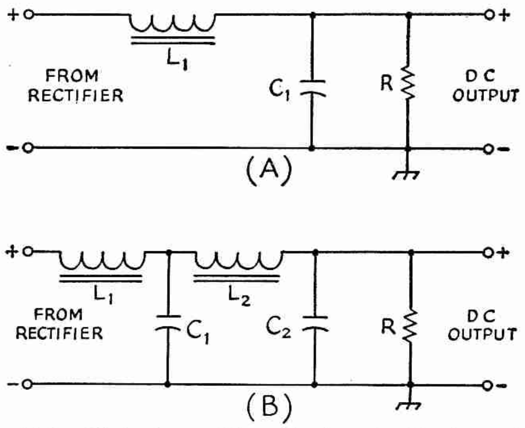


Fig. 7-9—Choke-input filter circuits. A—Single-section. B—Double-section.

choke (see Fig. 7-2) when the current drawn from the supply is small. This is in contrast to the capacitive-input filter in which the output voltage tends to soar toward the peak value of the rectified wave at light loads. Also, if the input choke has at least the critical value, the rectifier peak plate current will be limited to about twice the d.c. current drawn from the supply. Most rectifier tubes have peak-current ratings of three to four times their maximum d.c. output-current ratings. Therefore, with an input choke of at least critical inductance, current up to the maximum output-current rating of the rectifier may be drawn from the supply without exceeding the peak-current rating of the rectifier.

Minimum-Load—Bleeder Resistance

From the formula above for critical inductance, it is obvious that if no current is drawn from the supply, the critical inductance will be infinite. So that a practical value of inductance may be used, some current must be drawn from the supply at all times the supply is in use. From the formula we find that this minimum value of current is

$$I_{\text{MA.}} = \frac{E_{\text{VOLTS}}}{L_{\text{h}}}$$

Thus, if the choke has an inductance of 20 h., and the output voltage is 2000, the minimum load current should be 100 ma. This load may be provided, for example, by transmitter stages that draw current continuously (stages that are not keyed). However, in the majority of cases it will be most convenient to adjust the bleeder resistance so that the bleeder will draw the required minimum current. In the above example, the bleeder resistance should be 2000/0.1 = 20,000 ohms.

From the formula for critical inductance, it is seen that when more current is drawn from the supply, the critical inductance becomes less. Thus, as an example, when the total current, including the 100 ma. drawn by the bleeder rises to 400 ma., the choke need have an inductance of only 5 h. to maintain the critical value. This is fortunate, because chokes having the required inductance for the bleeder load only and that will maintain this value of inductance for much larger currents are very expensive.

Swinging Chokes

Less costly chokes are available that will maintain at least critical value of inductance over the range of current likely to be drawn from practical supplies. These chokes are called **swinging chokes.** As an example, a swinging choke may have an inductance rating of 5/25 h. and a current rating of 225 ma. If the supply delivers 1000 volts, the minimum load current should be 1000/25 = 40 ma. When the full load current of 225 ma. is drawn from the supply, the inductance will drop to 5 h. The critical inductance for 225 ma. at 1000 volts is 1000/225 = 4.5 h. Therefore the 5/25-h. choke maintains at least the critical inductance at the full current rating of 225 ma. At all load currents between 40 ma. and 225 ma., the choke will adjust its inductance to at least the approximate critical value.

Table 7-I shows the maximum supply output voltage that can be used with commonly-available swinging chokes to maintain critical inductance at the maximum current rating of the choke. These chokes will also maintain critical inductance for any *lower* values of voltage, or current down to the required minimum drawn by a proper bleeder as discussed above.

TABLE 7-I

L_h	*Max. ma.*	*Max. volts*	*Max.* R^1	*Min. ma.*^{2}
3.5/13.5	150	525	13.5K	39
5/25	175	875	25K	35
2/12	200	400	12K	33
5/25	200	1000	25K	40
5/25	225	1125	25K	45
2/12	250	500	12K	42
4/20	300	1200	20K	60
5/25	300	1500	25K	60
3/17	400	1200	17K	71
4/20	400	1600	20K	80
5/25	400	2000	25K	80
4/16	500	2000	16K	125
5/25	500	2500	25K	100
5/25	550	2750	25K	110

[1] Maximum bleeder resistance for critical inductance.
[2] Minimum current (bleeder) for critical inductance.

In the case of supplies for higher voltages in particular, the limitation on maximum load resistance may result in the wasting of an appreciable portion of the transformer power capacity in the bleeder resistance. Two input chokes in series will permit the use of a bleeder of twice the resistance, cutting the wasted current in half. Another alternative that can be used in a c.w. transmitter is to use a very high-resistance bleeder for protective purposes and only sufficient fixed bias on the tubes operating from the supply to bring the total current drawn from the supply, when the key is open, to the value of current that the required bleeder resistance should draw from the supply. Operating bias is brought back up to normal by increasing the grid-leak resistance. Thus the entire current capacity of the supply (with the exception of the small drain of the protective bleeder) can be used in operating the transmitter stages. With this system, it is advisable to operate the tubes at phone, rather than c.w., rating, since the average dissipation is increased.

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:

$$E_o = 0.9E_t - (I_B + I_L)(R_1 + R_2) - E_r$$

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 amperes; 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. The various voltage drops are shown in Fig. 7-12. 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.

Ripple with Choke Input

The percentage ripple output from a single-section filter (Fig. 7-9A) may be determined to a close approximation, for a ripple frequency of 120 cycles, from Fig. 7-10.

Example: L = 5 h., C = 4 μf., LC = 20.

From Fig. 7-10, percentage ripple = 5 per cent.

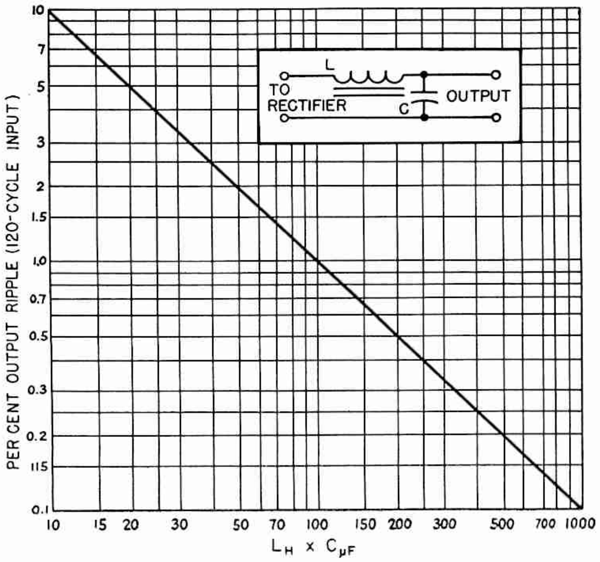


Fig. 7-10—Graph showing combinations of inductance and capacitance that may be used to reduce ripple with a single-section choke-input filter.

Example: L = 5 h. What capacitance is needed to reduce the ripple to 1 per cent? Following the 1-per-cent line to the right to its intersection with the diagonal, thence downward to the LC scale, read LC = 100. 100/5 = 20 μf.

In selecting values for the first filter section, the inductance of the choke should be determined by the considerations discussed previously. Then the capacitor should be selected that when combined with the choke inductance (minimum inductance in the case of a swinging choke) will bring the ripple down to the desired value. If it is found impossible to bring the ripple down to the desired figure with practical values in a single section, a second section can be added, as shown in Fig. 7-9B and the reduction factor from Fig. 7-8 applied as discussed under capacitive-input filters. The second choke should not be of the swinging type, but one having a more or less constant inductance with changes in current (smoothing choke).

OUTPUT CAPACITOR

If the supply is intended for use with an audio-frequency amplifier, the reactance of the last filter capacitor should be small (20 per cent or less) compared with the other audio-frequency 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 has a capacitance of 4 to 8 μf., 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 capacitor (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 capacitor 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. With a swinging choke, the minimum rated inductance of the choke should be used.

RATINGS OF FILTER COMPONENTS

Although filter capacitors in a choke-input filter are subjected to smaller variations in d.c. voltage than in the capacitive-input filter, it is advisable to use capacitors 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 capacitive-input type.

In a capacitive-input filter, the capacitors 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 capacitor voltage rating will be 550 × 1.41 or 775 volts. An 800-volt capacitor should be used, or preferably a 1000-volt unit.

Filter Capacitors in Series

Filter capacitors are made in several different types. Electrolytic capacitors, which are available for peak voltages up to about 800, combine high capacitance with small size, since the dielectric is an extremely thin film of oxide on aluminum foil. Capacitors of this type may be connected in series for higher voltages, although the filtering capacitance will be reduced to the resultant of the two capacitances in series. If this arrangement is used, it is important that *each* of the capacitors be shunted with a resistor of about 100 ohms per volt of supply voltage, with a power rating adequate for the total resistor current at that voltage. These resistors may serve as all or part of the bleeder resistance (see choke-input filters). Capacitors with higher-voltage ratings usually are made with a dielectric of thin paper impregnated with oil. The **working voltage** of a capacitor is the voltage that it will withstand continuously.

Filter Chokes

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 4 to 20 henrys ordinarily are used. When filter chokes are placed in the positive leads, the negative being grounded, the windings should be insulated from the core to withstand the full d.c. output voltage of the supply and be capable of handling the required load current.

Filter chokes or inductances are wound on iron cores, with a small gap in the core to prevent magnetic saturation of the iron at high currents. When the iron becomes saturated its permeability decreases, consequently the inductance also decreases. Despite the air gap, the inductance of a choke usually varies to some extent with the direct current flowing in the winding; hence it is necessary to specify the inductance at the current which the choke is intended to carry. Its inductance with little or no direct current flowing in the winding may be considerably higher than the value when full load current is flowing.

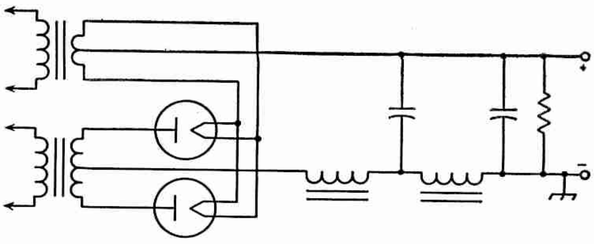


Fig. 7-11—In most applications, the filter chokes may be placed in the negative instead of the positive side of the circuit. This reduces the danger of a voltage breakdown between the choke winding and core.

● NEGATIVE-LEAD FILTERING

For many years it has been almost universal practice to place filter chokes in the positive leads of plate power supplies. This means that the insulation between the choke winding and its core (which should be grounded to chassis as a safety measure) must be adequate to withstand the output voltage of the supply. This voltage requirement is removed if the chokes are placed in the negative lead as shown in Fig. 7-11. With this connection, the capacitance of the transformer secondary to ground appears in parallel with the filter chokes tending to bypass the chokes. However, this effect will be negligible in practical application except in cases where the output ripple must be reduced to a very low figure. Such applications are usually limited to low-voltage devices such as receivers, speech amplifiers and v.f.o.'s where insulation is no problem and the chokes may be placed in the positive side in the conventional manner. In higher-voltage applications, there is no reason why the filter chokes should not be placed in the negative lead to reduce insulation requirements. Choke terminals, negative capacitor terminals and the transformer center-tap terminal should be well protected against accidental contact, since these will assume full supply voltage to chassis should a choke burn out or the chassis connection fail.

Plate and Filament Transformers

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:

$$E_t = 1.1\left[E_o + I(R_1 + R_2) + E_r\right]$$

where E_o is the required d.c. output voltage, I is the load current (including bleeder current) in amperes, 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

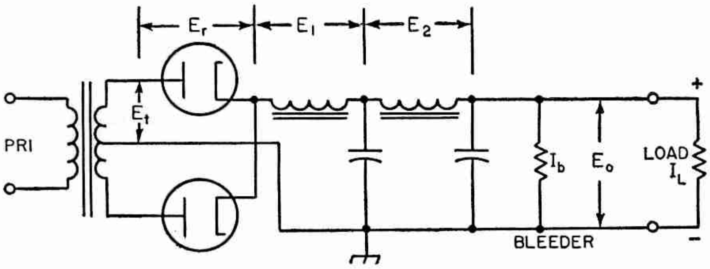


Fig. 7-12—Diagram showing various voltage drops that must be taken into consideration in determining the required transformer voltage to deliver the desired output voltage.

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 capacitive-input filter system can be calculated with Fig. 7-12.

Example:
Required d.c. output volts — 500
Load current to be drawn — 100 ma. (0.1 amp)

$$\text{Load resistance} = \frac{500}{0.1} = 5000 \text{ ohms.}$$

If the rectifier resistance is 200 ohms, Fig. 7-5 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

$$E_t = \frac{E_o + I\left(R_1 + R_2 + R_r\right)}{1.15}$$

$$= \frac{500 + 0.1\left(200 + 300 + 200\right)}{1.15}$$

$$= \frac{570}{1.15} = 495 \text{ volts.}$$

Volt-Ampere Rating

The volt-ampere rating of the transformer depends upon the type of filter (capacitive or choke input). With a capacitive-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.

Broadcast & Television Replacement Transformers in Amateur Transmitter Service

Small power transformers of the type sold for replacement in broadcast and television receivers are usually designed for service in terms of use for several hours continuously with capacitor-input filters. In the usual type of amateur transmitter service, where most of the power is drawn intermittently for periods of several minutes with equivalent intervals in between, the published ratings can be exceeded without excessive transformer heating.

With capacitor input, it should be safe to draw 20 to 30 per cent more current than the rated value. With a choke-input filter, an increase in current of about 50 per cent is permissible. If a bridge rectifier is used (with a choke-input filter) the output voltage will be approximately doubled. In this case, it should be possible in amateur transmitter service to draw the rated current, thus obtaining about twice the rated output power from the transformer.

This does not apply, of course, to amateur transmitter plate transformers which are usually already rated for intermittent service.

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.

Typical Power Supplies

Figs. 7-13 and 7-14 show typical power-supply circuits. Fig. 7-13 is for use with transformers commonly listed as broadcast or television replacement power transformers. In addi-

Typical Power Supplies

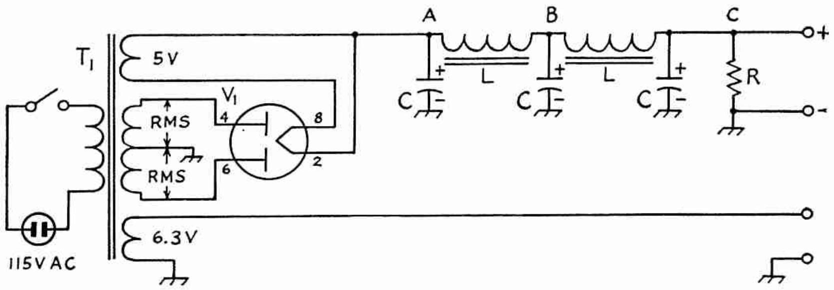


Fig. 7-13—Typical a.c. power-supply circuit for receivers, exciters, or low-power transmitters. Representative values will be found in Table 7-II. The 5-volt winding of T_1 should have a current rating of at least 2 amp. for types 5Y3-GT and 5V4-GA, and 3 amp. for 5U4-GB.

tion to the high-voltage winding for plate supply, these transformers have windings that supply filament voltages for both the rectifier tube and the 6.3-volt tubes in the receiver or low-power transmitter or exciter. Transformers of this type may be obtained in ratings up to 1200 volts r.m.s. center-tapped, 200 d.c. ma. output.

Fig. 7-13 shows a two-section filter with capacitor input. However, depending upon the maximum hum level that may be allowable for a particular application, the last capacitor and choke may not be needed. In some low-current applications, the first capacitor alone may provide adequate filtering. Table 7-II shows the approximate full-load and bleeder-load output voltages and a.c. ripple percentages for several representative sets of components. Voltage and ripple values are given for three points in the circuit — Point A (first capacitor only used), Point B (last capacitor and choke omitted), and Point C (complete two-section filter in use). In each case, the bleeder resistor *R* should be used across the output.

Table 7-II also shows approximate output voltages and ripple percentages for choke-input filters (first filter capacitor omitted), for Point B (last capacitor and choke omitted), and Point C (complete two-section filter, first capacitor omitted).

Actual full-load output voltages may be somewhat lower than those shown in the table, since the voltage drop through the resistance of the transformer secondary has not been included.

Fig. 7-14 shows the conventional circuit of a transmitter plate supply for higher powers. A full-wave rectifier circuit, half-wave rectifier tubes, and separate transformers for high voltage, rectifier filaments and transmitter filaments are used. The high-voltage transformers used in this circuit are usually rated directly in terms of d.c. output voltage, assuming rectifiers and filters of the type shown in Fig. 7-14. Table 7-III shows typical values for representative supplies, based on commonly available components. Transformer

TABLE 7-II

Capacitor-Input Power Supplies

T_1 Rating		*V_1 Tube Type*	*C*		*L*		*R*		*Approximate Full-load d.c. Volts at*			*Approximate Ripple % at*			*Approx. Output Volts Bleeder Load*	*Useful Output Ma.**
Volts R.M.S. (C.T.)	*Ma. D.C.*		*µf.*	*Volts*	*H.*	*Ohms*	*Ohms*	*Watts*	*A*	*B*	*C*	*A*	*B*	*C*		
650	40	5Y3-GT	8	600	8	400	90K	5	375	360	345	2.5	0.08	0.002	450	36
650	40	5V4-GA	8	600	8	400	90K	5	410	395	375	2.5	0.08	0.002	450	36
700	90	5Y3-GT	8	600	10	225	46K	10	370	350	330	6	0.1	0.002	460	82
700	90	5V4-GA	8	600	10	225	46K	10	410	390	370	6	0.1	0.002	460	82
750	150	5U4-GB	8	700	8	145	25K	10	375	350	330	9	0.2	0.006	500	136
750	150	5V4-GA	8	700	8	145	25K	10	425	400	380	9	0.2	0.006	500	136
800	200	5U4-GB	8	700	8	120	22K	20	375	350	325	12	0.3	0.008	550	184
Choke-Input Power Supplies																
650	40	5Y3-GT	8	450	15	420	18K	10	—	240	225	—	0.8	0.01	265	25
650	40	5V4-GA	8	450	15	420	18K	10	—	255	240	—	0.8	0.01	280	25
700	90	5Y3-GT	8	450	10	225	11K	10	—	240	220	—	1.25	0.02	250	68
700	90	5V4-GA	8	450	10	225	11K	10	—	270	250	—	1.25	0.02	280	68
750	150	5Y3-GT	8	450	12	150	13K	20	—	265	245	—	1	0.015	325	125
750	150	5V4-GA	8	450	12	150	13K	20	—	280	260	—	1	0.015	340	125
800	200	5U4-GB	8	450	12	140	14K	20	—	275	250	—	1	0.015	350	175

* Balance of transformer current capacity consumed by bleeder resistor.

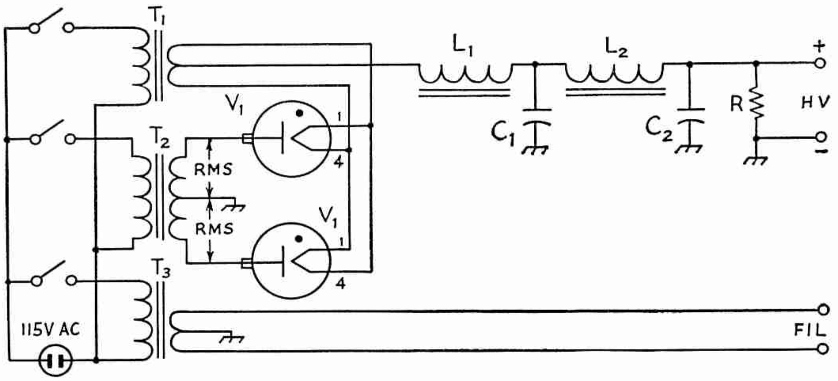


Fig. 7-14—Conventional power-supply circuit for higher-power transmitters.

C_1, C_2—4 μf. for approximately 0.5% output ripple; 2 μf. for approximately 1.5% output ripple. C_2 should be 4 μf. if supply is for modulator.

R—25,000 ohms.

L_1—Swinging choke: 5/25 h., current rating same as T_2.

L_2—Smoothing choke: current rating same as T_2.

T_1—2.5 volts, 4 amp. for type 816; 2.5 volts, 10 amp. for 866A.

T_2—D.c. voltage rating same as output voltage.

T_3—Voltage and current rating to suit transmitter-tube requirements.

V_1—Type 816 for 400/500-volt supply; 866A for others shown in Table 7-III.

See Table 7-III for other values.

voltages shown are representative for units with dual-voltage secondaries. The bleeder-load voltages shown may be somewhat lower than actually found in practice, because transformer resistance has not been included. Ripple at the output of the first filter section will be approximately 5 per cent with a 4-μf. capacitor, or 10 per cent with a 2-μf. capacitor. Transformers made for amateur service are designed for choke-input. If a capacitor-input is used rating should be reduced about 30%.

TABLE 7-III

Approx. D.C. Output		*T_2 Rating*		*L_2 H.*	*Voltage Rating C_1, C_2*	*R Watts*	*Approx. Bleeder-Load Output Volts*
Volts	*Ma.*[1]	*Approx. V.R.M.S.*	*Ma.*				
400/500	230	520/615	250	4	700	20	440/540
600/750	260	750/950	300	8	1000	50	650/800
1250/1500	240	1500/1750	300	8	2000	150	1300/1600
1250/1500	440	1500/1750	500	6	2000	150	1315/1615
2000/2500	200	2400/2900	300[4]	8	3000	320[2]	2050/2550
2000/2500	400	2400/2900	500	6	3000	320[2]	2065/2565
2500/3000	380	2500/3450	500[5]	6	4000	500[3]	2565/3065

[1] Balance of transformer current rating consumed by bleeder resistor.
[2] Use two 160-watt, 12,500-ohm units in series.
[3] Use five 100-watt, 5000-ohm units in series.
[4] Regulation will be somewhat better with a 400- or 500-ma. choke.
[5] Regulation will be somewhat better with a 550-ma. choke.

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 an available power supply. 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. 7-15A. 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 20-watt resistor is the nearest safe rating to be used.

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 im-

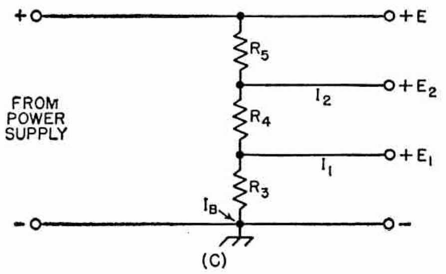


Fig. 7-15—A—Series voltage-dropping resistor. B—Simple voltage divider. C—Multiple divider circuit.

$$R_3 = \frac{E_1}{I_b};\ R_4 = \frac{E_2 - E_1}{I_b + I_1};\ R_5 = \frac{E - E_2}{I_b + I_1 + I_2}$$

proved somewhat by connecting a second resistor from the low-voltage end of the first to the negative power-supply terminal, as shown in Fig. 7-15B. 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. 7-15C. 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 the caption of Fig. 7-15C, 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 needed 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 (0C3/VR105, 0D3/VR150, 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 near 150, 105, 90 and 75 volts.

The fundamental circuit for a gaseous regulator is shown in Fig. 7-16A. 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 to 40 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.

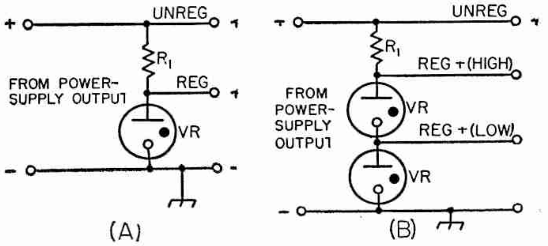


Fig. 7-16—Voltage-stabilizing circuits using VR tubes.

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:

$$R = \frac{(E_s - E_r)}{I}$$

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

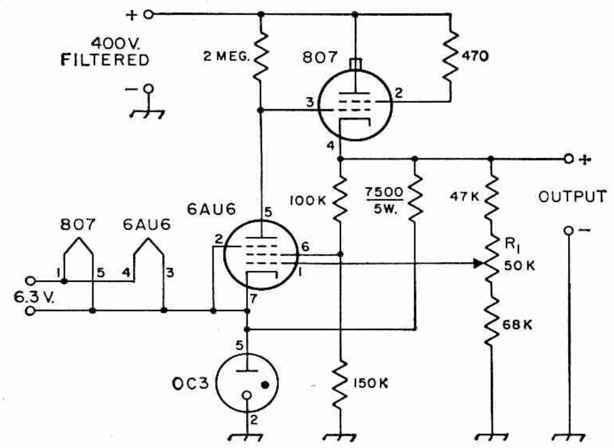


Fig. 7-17—Electronic voltage-regulator circuit. Resistors are ½ watt unless specified otherwise.

I is the maximum tube current in amperes, (usually 40 ma., or 0.04 amp.).

Fig. 7-16B 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 these regulator circuits.

A single VR tube may also be used to regulate the voltage to a load current of almost any value so long as the *variation* in the current does not exceed 30 to 35 ma. If, for example, the average load current is 100 ma., a VR tube may be used to hold the voltage constant provided the current does not fall below 85 ma. or rise above 115 ma. In this case, the resistance should be calculated to drop the voltage to the VR-tube rating at the maximum load current to be expected plus about 5 ma. If the load resistance is constant, the effects of variations in line voltage may be eliminated by basing the resistance on the load current plus 15 ma. Voltage-regulator tubes may also be connected in parallel as described later in this chapter.

Electronic Voltage Regulation

Several circuits have been developed for regulating the voltage output of a power supply elec-

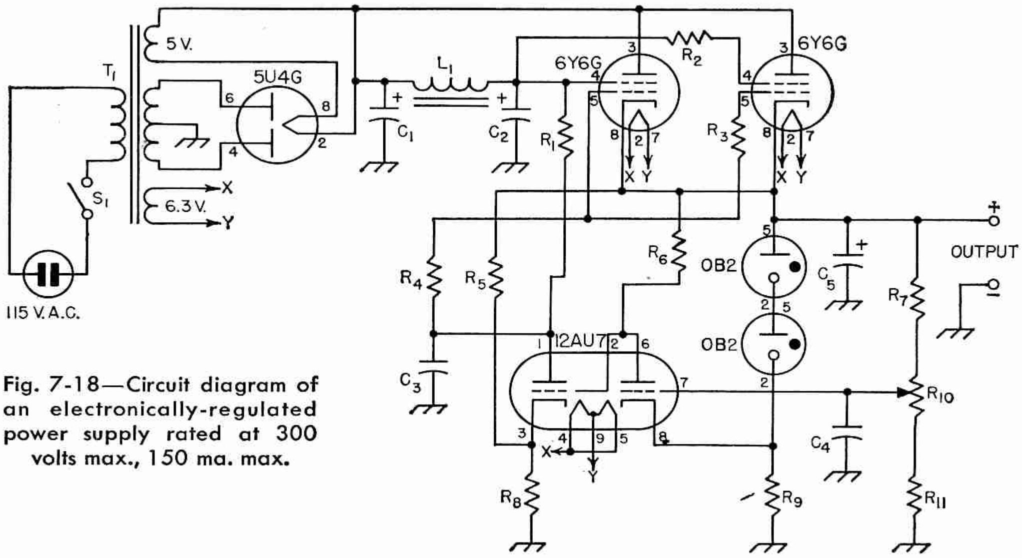


Fig. 7-18—Circuit diagram of an electronically-regulated power supply rated at 300 volts max., 150 ma. max.

C_1, C_2, C_5—16-μf. 600-volt electrolytic.
C_3—0.015-μf. paper.
C_4—0.1-μf. paper.
R_1—0.3 megohm, ½ watt.
R_2, R_3—100 ohms, ½ watt.
R_4—510 ohms, ½ watt.
R_5, R_8—30,000 ohms, 2 watts.
R_6—0.24 megohm, ½ watt.
R_7—0.15 megohm, ½ watt.
R_9—9100 ohms, 1 watt.
R_{10}—0.1-megohm potentiometer.
R_{11}—43,000 ohms, ½ watt.
L_1—8-hy., 40-ma. filter choke.
S_1—S.p.s.t. toggle.
T_1—Power transformer: 375-375 voltsr.m.s., 160 ma.; 6.3 volts, 3 amps.; 5 volts, 3 amps. (Thor. 22R33).

tronically. While more complicated than the VR-tube circuits, they will handle higher voltages and currents and the output voltage may be varied continuously over a wide range. In the circuit of Fig. 7-17, the 0C3 regulator tube supplies a reference of approximately 105 volts for the 6AU6 control tube. When the load connected across the output terminals increases, the output voltage tends to decrease. This drops the voltage on the control grid of the 6AU6, causing the tube to draw less current through the 2-megohm plate resistor. As a consequence the grid voltage on the 807 series regulator rises and the voltage drop across the 807 decreases, compensating for the reduction in output voltage. With the values shown, adjustment of R_1 will give a regulated output from 150 to 250 volts, at up to 60 or 70 ma. A 6L6-GB can be substituted for the type 807; the available output current can be increased by adding one or more tubes in parallel with the series regulator tube. When tubes are connected in parallel, 100-ohm resistors should be wired to each control grid and plate terminal, to reduce the chances for parasitic oscillations.

Another similar regulator circuit is shown in Fig. 7-18. The principal difference is that screen-grid regulator tubes are used. The fact that a screen-grid tube is relatively insensitive to changes in plate voltage makes it possible to obtain a reduction in ripple voltage adequate for many purposes simply by supplying filtered d.c. to the screens with a consequent saving in weight and cost. The accompanying table shows the performance of the circuit of Fig. 7-18. Column I shows various output voltages, while Column II shows the maximum current that can be drawn at that voltage with negligible variation in output voltage. Column III shows the measured ripple at the maximum current. The second part of the table shows the variation in ripple with load current at 300 volts output.

Table of Performance for Circuit of Fig. 7-18

I	*II*	*III*	*Output voltage — 300*	
450 v.	22 ma.	3 mv.	150 ma.	2.3 mv.
425 v.	45 ma.	4 mv.	125 ma.	2.8 mv.
400 v.	72 ma.	6 mv.	100 ma.	2.6 mv.
375 v.	97 ma.	8 mv.	75 ma.	2.5 mv.
350 v.	122 ma.	9.5 mv.	50 ma.	3.0 mv.
325 v.	150 ma.	3 mv.	25 ma.	3.0 mv.
300 v.	150 ma.	2.3 mv.	10 ma.	2.5 mv.

High-Voltage Regulators

Regulated screen voltage is required for screen-grid tubes used as linear amplifiers in single-side-band operation. Figs. 7-19 through 7-22 show various different circuits for supplying regulated voltages up to 1200 volts or more.

In the circuit of Fig. 7-19, gas-filled regulator tubes are used to establish a fixed reference voltage to which is added an electronically regulated variable voltage. The design can be modified to give any voltage from 225 volts to 1200 volts, with each design-center voltage variable by plus or minus 60 volts.

The output voltage will depend upon the number and voltage ratings of the VR tubes in the string between the 991 and ground. The total VR-tube voltage rating needed can be determined by subtracting 250 volts from the desired output voltage. As examples, if the desired output voltage is 350, the total VR-tube voltage rating should be 350 − 250 = 100 volts. In this case, a VR-105 would be used. For an output voltage of 1000, the VR-tube voltage rating should be 1000 − 250 = 750 volts. In this case, five VR-150s would be used in series.

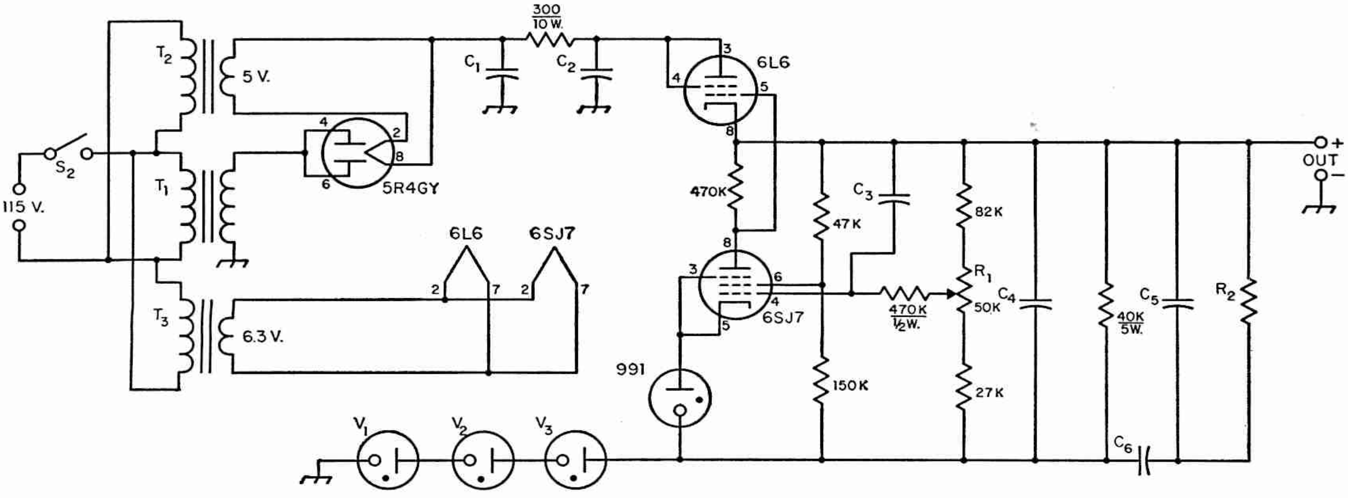


Fig. 7-19—High-voltage regulator circuit by W4PRM. Resistors are 1 watt unless indicated otherwise.

C_1, C_2—4-μf. paper, voltage rating above peak-voltage output of T_1.
C_3—0.1-μf. paper, 600 volts.
C_4—12-μf. electrolytic, 450 volts.
C_5—40 μf., voltage rating above d.c. output voltage. Can be made up of a combination of electrolytics in series, with equalizing resistor. (See section on ratings of filter components.)
C_6—4-μf. paper, voltage rating above voltage rating of VR string.
R_1—50,000-ohm, 4-watt potentiometer.
R_2—Bleeder resistor, 50,000 to 100,000 ohms, 25 watts (not needed if equalizing resistors mentioned above are used).
T_1—See text.
T_2—Filament transformer; 5 volts, 2 amp.
T_3—Filament transformer; 6.3 volts, 1.2 amp.
V_1, V_2, V_3—See text.

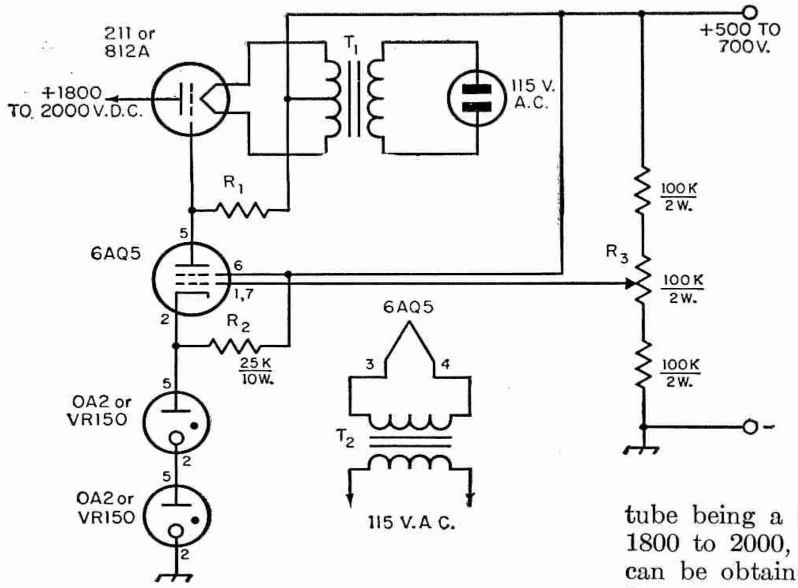


Fig. 7-20—Screen regulator circuit designed by W9OKA. Resistances are in ohms (K = 1000).
R_1—6000 ohms for 211; 2300 ohms for 812A, 20 watts.
R_2—25,000 ohms, 10 watts.
R_3—Output voltage control, 0.1-megohm, 2-watt potentiometer.
T_1—Filament transformer: 10 volts, 3.25 amp. for 211; 6.3 volts, 4 amp. for 812A.
T_2—Filament transformer: 6.3 volts, 1 amp.

The maximum voltage output that can be obtained is approximately equal to 0.7 times the r.m.s. voltage of the transformer T_1. The current rating of the transformer must be somewhat above the load current to take care of the voltage dividers and bleeder resistances.

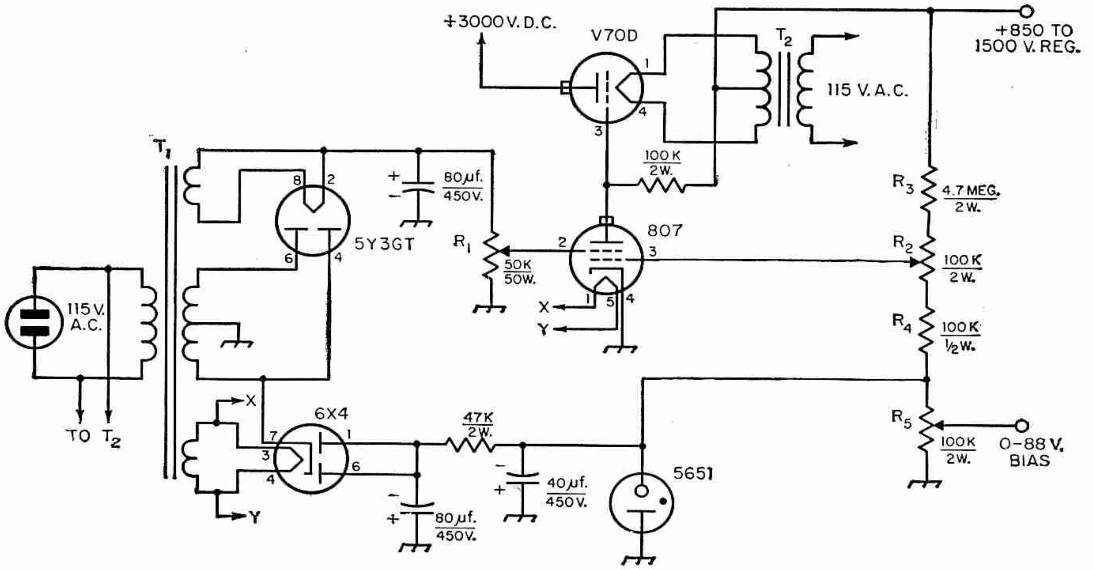
A single 6L6 will handle 90 ma. For larger currents, 6L6s may be added in parallel.

The heater circuit supplying the 6L6 and 6SJ7 should *not* be grounded. The shaft of R_1 should be grounded. When the output voltage is above 300 or 400, the potentiometer should be provided with an insulating mounting, and should be controlled from the panel by an extension shaft with an insulated coupling and grounded control.

In some cases where the plate transformer has sufficient current-handling capacity, it may be desirable to operate a screen regulator from the plate supply, rather than from a separate supply. This can be done if a regulator tube is used that can take the required voltage drop. In Fig. 7-20, a type 211 or 812A is used, the control tube being a 6AQ5. With an input voltage of 1800 to 2000, an output voltage of 500 to 700 can be obtained with a regulation better than 1 per cent over a current range of 0 to 100 ma.

In the circuit of Fig. 7-21, a V-70D (or 8005) is used as the regulator, and the control tube is an 807 which can take the full output voltage, making it unnecessary to raise it above ground with VR tubes. If taps are switched on R_1, the output voltage can be varied over a wide range. Increasing the screen voltage decreases the output voltage. For each position of the tap on R_1, decreasing the value of R_3 will lower the minimum output voltage as R_2 is varied, and decreasing the

Fig. 7-21—This regulator circuit used by W1SUN operates from the plate supply and requires no VR string. A small supply provides screen voltage and reference bias for the control tube.
Unless otherwise marked, resistances are in ohms. (K = 1000). Capacitors are electrolytic.
R_1—50,000-ohm, 50-watt adjustable resistor.
R_2—0.1-megohm 2-watt potentiometer.
R_3—4.7 megohms, 2 watts.
R_4—0.1 megohm, ½ watt.
T_1—Power transformer: 470 volts center tapped, 40 ma.; 5 volts, 2 amps.; 6.3 volts, 2 amps.
T_2—Filament transformer: 7.5 volts, 3.25 amp. (for V-70D).



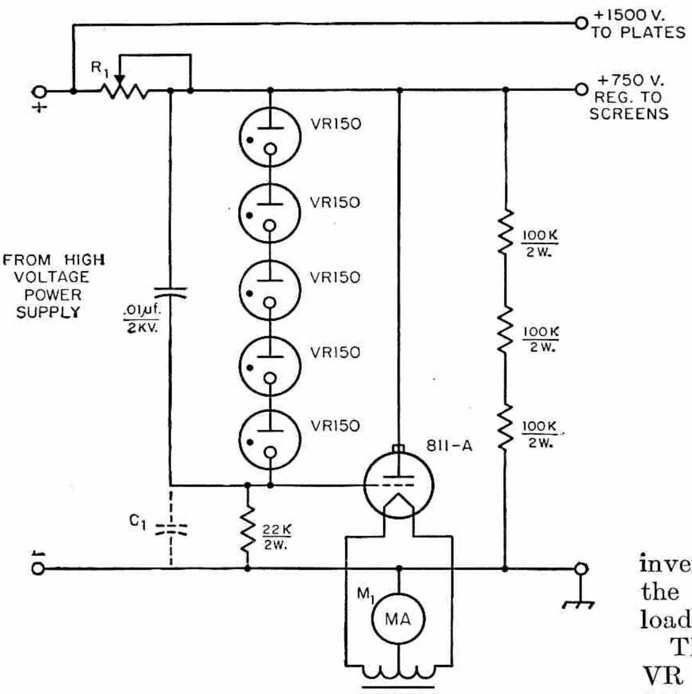


Fig. 7-22—Shunt screen regulator used by W2AZW. Resistances are in ohms (K = 1000).

C1—0.01 μf., 400 volts if needed to suppress oscillation.

M1—See text.

R1—Adjustable wire-wound resistor, resistance and wattage as required.

value of R_4 will raise the maximum output voltage. However, if these values are made too small, the 807 will lose control.

At 850 volts output, the variation over a current change of 20 to 80 ma. should be negligible. At 1500 volts output with the same current change, the variation in output voltage should be less than three per cent. Up to 88 volts of grid bias for a Class A or Class AB_1 amplifier may be taken from the potentiometer across the reference-voltage source. This bias cannot, of course, be used for biasing a stage that is drawing grid current.

A somewhat different type of regulator is the shunt regulator shown in Fig. 7–22. The VR tubes and R_2 in series are across the output. Since the voltage drop across the VR tubes is constant, any change in output voltage appears across R_2. This causes a change in grid bias on the 811-A grid, causing it to draw more or less current in inverse proportion to the current being drawn by the amplifier screen. This provides a constant load for the series resistor R_1.

The output voltage is equal to the sum of the VR drops plus the grid-to-ground voltage of the 811-A. This varies from 5 to 20 volts between full load and no load. The initial adjustment is made by placing a milliammeter in the filament center-tap lead, as shown, and adjusting R_1 for a reading of 15 to 20 ma. higher than the normal peak screen current. This adjustment should be made with the amplifier connected but with no excitation, so that the amplifier draws idling current. After the adjustment is complete, the meter may be removed from the circuit and the filament center tap connected directly to ground. Adjustment of the tap on R_1 should, of course, be made with the high voltage turned off.

Any number of VR tubes may be used to provide a regulated voltage near the desired value· The maximum current through the 811-A should be limited to the maximum plate-current rating of the tube. If larger currents are necessary, two 811-As may be connected in parallel. Over a current range of 5 to 60 ma., the regulator holds the output voltage constant within 10 or 15 volts.

Bias Supplies

As discussed in Chapter 6 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. 7-23A 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 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.

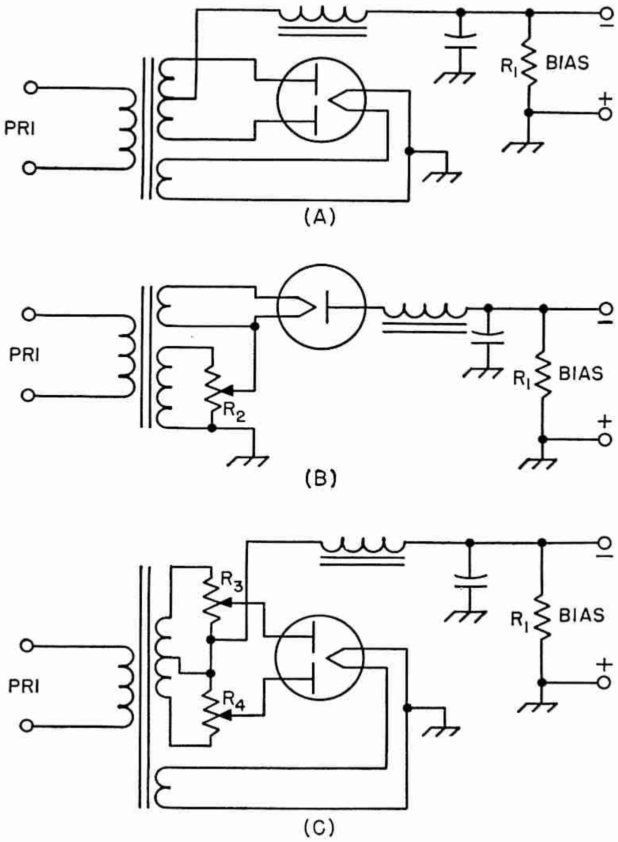


Fig. 7-23—Simple bias-supply circuits. In A, the peak transformer voltage must not exceed the operating value of bias. The circuits of B (half-wave) and C (full-wave) may be used to reduce transformer voltage to the rectifier. R_1 is the recommended grid-leak resistance.

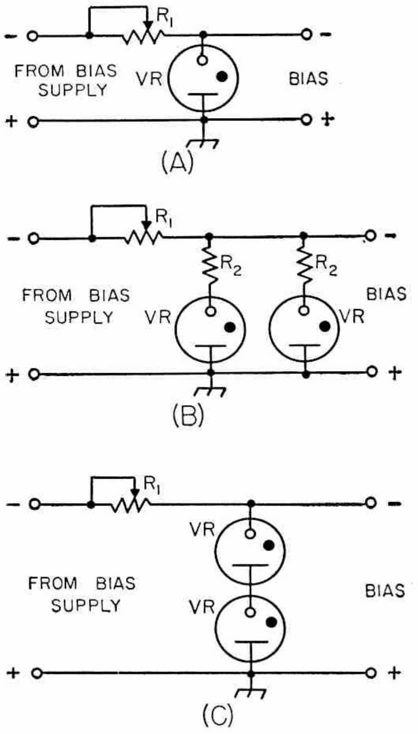


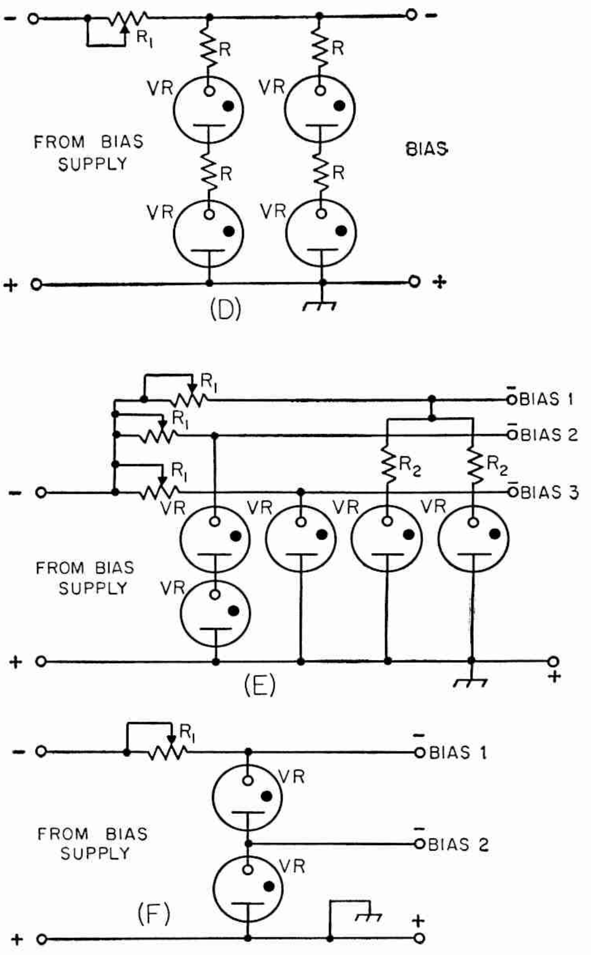
Fig. 7-24—Illustrating the use of VR tubes in stabilizing protective-bias supplies. R_1 is a resistor whose value is adjusted to limit the current through each VR tube to 5 ma. before amplifier excitation is applied. R and R_2 are current-equalizing resistors of 50 to 1000 ohms.

A full-wave circuit is shown in Fig. 7-23C. R_3 and R_4 should have the same total resistance and the taps should be adjusted symmetrically. In all cases, the transformer must be designed to furnish the current drawn by these resistors plus the current drawn by R_1.

Regulated Bias Supplies

The inconvenience of the circuits shown in Fig. 7-23 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. 7-24A. 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 resistor, as discussed in the 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. 7-24B, for each 40 ma., or less, of additional grid current. The



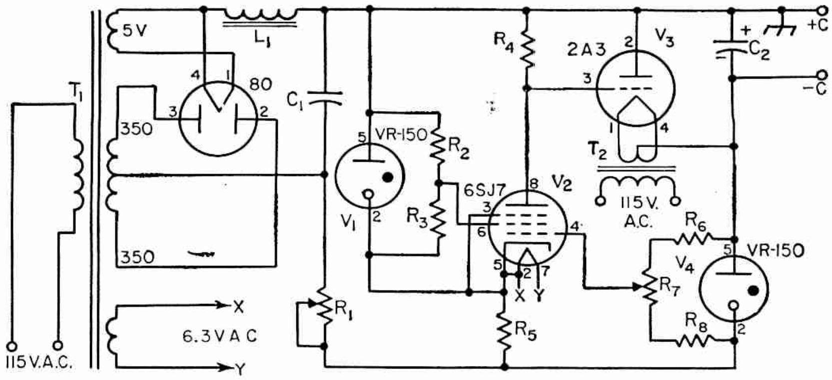


Fig. 7-25—Circuit diagram of an electronically-regulated bias supply.

C_1—20-μf. 450-volt electrolytic.
C_2—20-μf. 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.
T_2—2.5-volt filament transformer (Thordarson 21F00).

resistors R_2 are for the purpose of helping to maintain equal currents through each VR tube, and should have a value of 50 to 1000 ohms or more.

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 the diagrams of Fig. 7-24C 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 in Fig. 7-24E, to adapt them to 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. 7-25. The output voltage may be adjusted to any value between 40 volts and 80 volts and the unit will handle grid currents up to 35 ma. over the range of 50 to 80 volts, and 25 ma. over the remainder of the range. If higher current-handling capacity is required, more 2A3s can be connected in parallel with V_3. The regulation will hold to about 0.01 volt per milliampere of grid current. The regulator operates as follows: Since the voltage drop across V_3 and V_4 is in parallel with the voltage drop across V_1 and R_5, any change in voltage across V_3 will appear across R_5 because the voltage drops across both VR tubes remain constant. R_5 is a cathode biasing resistor for V_2, so any voltage change across it appears as a grid-voltage change on V_2. This change in grid voltage is amplified by V_2 and appears across R_4 which is connected to the plate of V_2 and the grids of V_3. This change in voltage swings the grids of V_3 more positive or negative, and thus varies the internal resistance of V_3, maintaining the voltage drop across V_3 practically constant.

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. 7-26A. In another arrangement, shown at B, a spare filament winding can be used to operate a filament transformer of similar voltage rating in reverse to obtain a voltage of about 130 from the winding that is customarily the primary. This

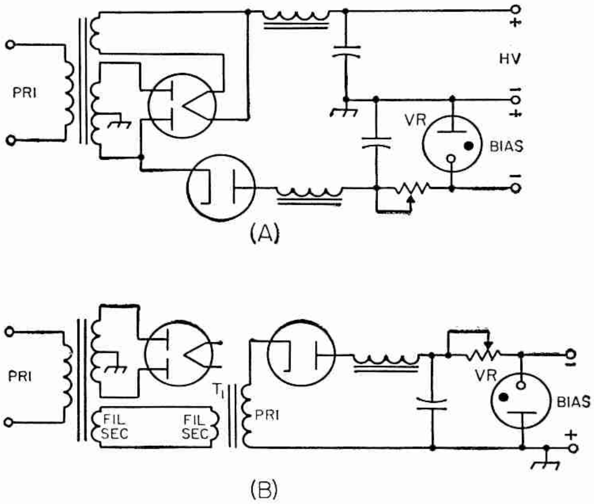


Fig. 7-26—Convenient means of obtaining biasing voltage. A—From a low-voltage plate supply. B—From spare filament winding. T_1 is a filament transformer, of a voltage output similar to that of the spare filament winding, connected in reverse to give 115 volts r.m.s. output. If cold-cathode or selenium rectifiers are used, no additional filament supply is required.

will be sufficient to operate a VR75 or VR90 regulator tube.

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. 7-27, 7-28 and 7-29 may be used with any type of rectifier, they find their greatest advantage when used with selenium rectifiers which require no filament transformer. These circuits must be used with caution, observing line polarity in the circuits so marked, to avoid shorting the line, since the negative output terminal should always be grounded. In circuits showing isolating transformers, the transformer is a requirement, since without the transformer, the negative output terminal cannot be grounded in following good practice for safety without shorting out part of the rectifier circuit. In the circuits which do not show a transformer, the transformer is preferable, since it avoids the need for a correctly polarized power-line connection to prevent a short circuit.

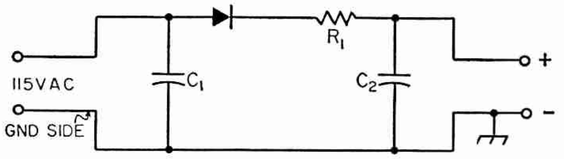


Fig. 7-27—Simple half-wave circuit for selenium rectifier.

C$_1$—0.05-μf. 600-volt paper.
C$_2$—40-μf. 200-volt electrolytic.
R$_1$—25 to 100 ohms.

Fig. 7-27 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 as a bias supply by reversing the polarity of the rectifier and capacitors.

Three voltage-doubler circuits are shown in Fig. 7-28. At A is a full-wave circuit, while the other two, at A and B, are half-wave circuits. Although easier to filter, the circuit of A has the disadvantage that the output cannot be grounded directly unless an isolation transformer is used. B and C are similar, except that the series capacitor is in different sides of the circuit. The output of B can be grounded directly if proper line polarity is observed. Circuit C, which includes a filter for illustration purposes, requires an isolation transformer if the output is to be grounded, but since all three capacitors, including the filter capacitor, have a common negative connection, a triple-unit capacitor may be used where space must be conserved.

Fig. 7-29 shows voltage tripler and quadrupler circuits. The circuit of A is a halfwave tripler. A full-wave tripler, requiring an additional rectifier element, is shown at B. The circuit of Fig. 7-29C is a half-wave voltage quadrupler. The full-wave version is shown at D. Both full-wave circuits require an isolation transformer to permit grounding of the output.

In the circuits of Figs. 7-28 and 7-29 where an isolation transformer is not shown, it is essential that the indicated line polarity be observed if the output is to be grounded. Otherwise part of the circuit will be shorted out.

The resistors R_1 are for rectifier protective purposes, and recommended minimum values are given in the table at the end of this section. The value of capacitance given is representative. Larger values will improve voltage regulation. Smaller values may be used at a sacrifice in regulation.

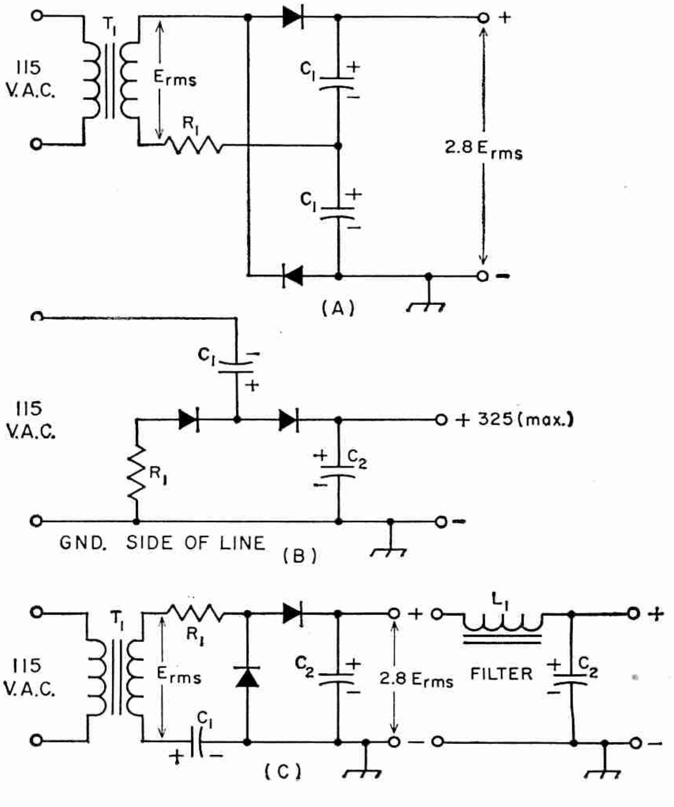


Fig. 7-28—Voltage-doubling circuits for use with selenium rectifiers. Maximum back voltage on rectifiers is 2.8 E_{rms}. Voltage rating at least 1.4 E_{rms} for C$_1$, at least 2.8 E_{rms} for C$_2$.

C$_1$—40-μf. electrolytic.
C$_2$—40-μf. electrolytic.
R$_1$—25 to 100 ohms.
L$_1$—Filter choke.
T$_1$—Isolation transformer.

Selenium Rectifiers

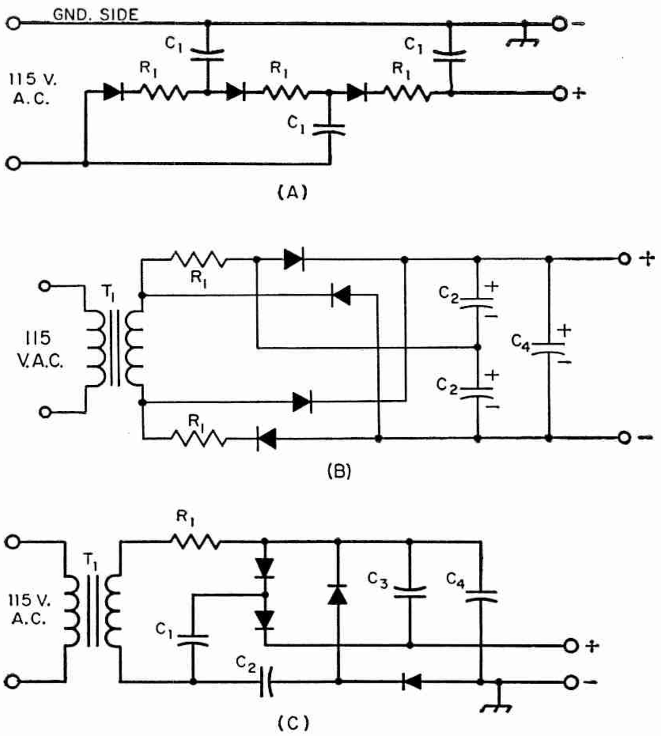


Fig. 7-29—A—Tripler circuit. B—Half-wave quadrupler. C—Full-wave quadrupler.
C_1—40-μf. 200-volt electrolytic.
C_2—40-μf. 450-volt electrolytic.
C_3—48-μf. 600-volt electrolytic (three 16-μf. units in parallel).
C_4—48-μf. 700-volt electrolytic (three 16-μf. units in parallel).
R_1—25 to 100 ohms.
T_1—Isolating transformer.

Power-Line Considerations

● 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. 7-30A. 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. 7-30B. Furthermore, with the neutral open, the voltage will then be divided between the two sides in inverse 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

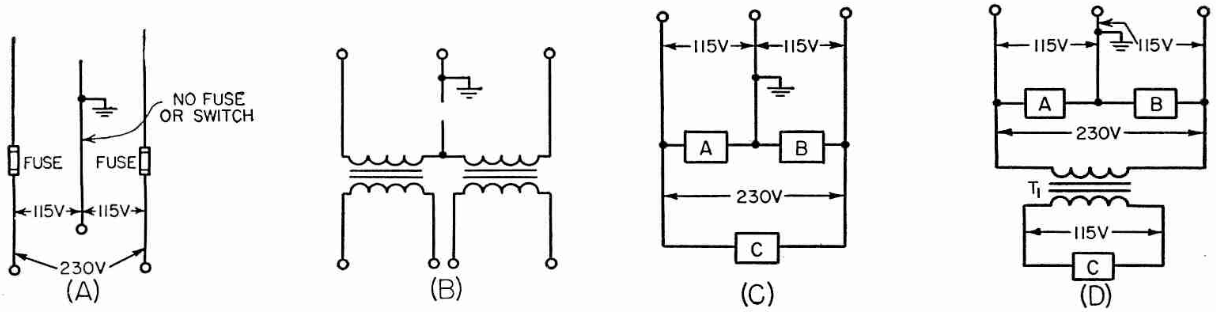


Fig. 7-30—Three-wire power-line circuits. A—Normal 3-wire-line termination. No fuse should be used in the grounded (neutral) line. B—Showing that a switch in the neutral does not remove voltage from ether side of the line. C—Connections for both 115- and 230-volt transformers. D—Operating a 115-volt plate transformer from the 230-volt line to avoid light blinking. T_1 is a 2-to-1 step-down transformer.

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 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 load can be distributed to keep the line balanced. 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 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. 7-30C. 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. 7-30D).

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.

Fusing

All transformer primary circuits should be properly fused. To determine the approximate current rating of the fuse to be used, multiply each current being drawn from the supply in amperes by the voltage at which the current is being drawn. Include the current taken by bleeder resistances and voltage dividers. In the case of series resistors, use the source voltage, not the voltage at the equipment end of the resistor. Include filament power if the transformer is supplying filaments. After multiplying the various voltages and currents, add the individual products. Then divide by the line voltage and add 10 or 20 per cent. Use a fuse with the nearest larger current rating.

● 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. 7-31A. A toy transformer is used to boost or buck the line voltage as required. The transformer should have a tapped secondary varying between 6 and 20 volts in steps of 2 or 3 volts and its secondary should be capable of carrying the full load current of the entire transmitter, or that portion of it fed by the toy transformer.

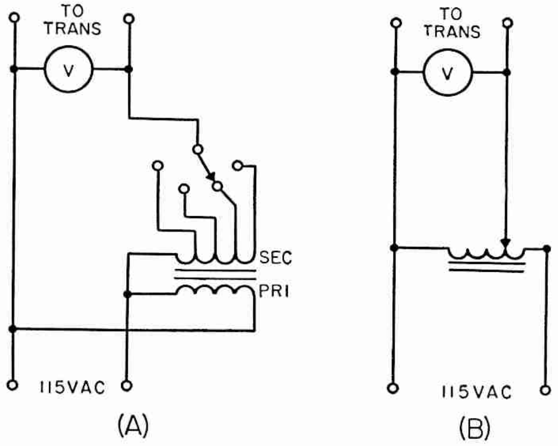


Fig. 7-31—Two methods of transformer primary control. At A is a tapped toy transformer which may be connected so as to boost or buck the line voltage as required. At B is indicated a variable transformer or autotransformer (Variac) which feeds the transformer primaries.

The secondary is connected in series with the line voltage and, if the phasing of the windings is correct, the voltage applied to the primaries of the transmitter transformers can be brought up to the rated 115 volts by setting the toy-transformer tap switch on the right tap. If the phasing of the two windings of the toy transformer happens to be reversed, the voltage will be reduced instead of increased. This connection may be used in cases where the line voltage may be above 115 volts. This method is preferable to using a resistor in the primary of a power transformer since it does not affect the voltage regulation as seriously. The circuit of 7-31B illustrates the use of a variable autotransformer (Variac) for adjusting line voltage.

Another scheme by which the primary voltage of each transformer in the transmitter may be adjusted to give a desired secondary voltage, with a master control for compensating for changes in line voltage, is shown in Fig. 7-32.

This arrangement has the following features:

1) Adjustment of the switch S_1 to make the voltmeter read 105 volts automatically adjusts all transformer primaries to the predetermined correct voltage.

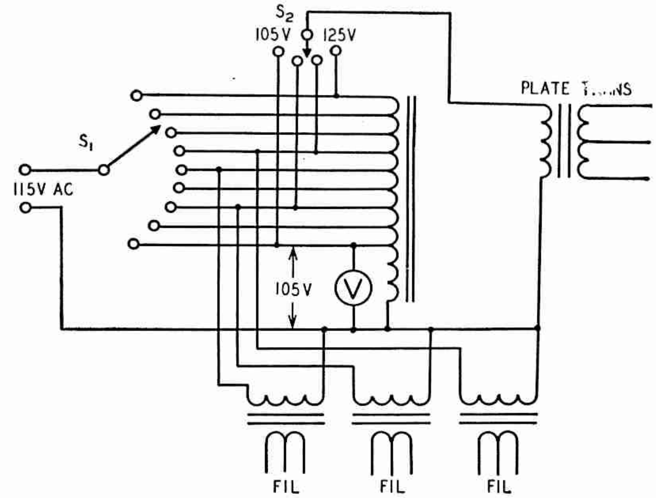


Fig. 7-32—With this circuit, a single adjustment of the tap switch S_1 places the correct primary voltage on all transformers in the transmitter.

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., as required to obtain the desired output voltage.

3) Independent control of the plate transformer is afforded by the tap switch S_2. This permits power-input control and does not require an extra autotransformer.

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 v.a. 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. All negative terminals of plate supplies and positive terminals of bias supplies should be securely grounded to the chassis, and the chassis connected to a waterpipe or radiator ground. All transformer, choke, and capacitor cases should also be grounded to the chassis. A.c. power cords and chassis connectors should be arranged so that exposed contacts are never "live." Starting at the conventional a.c. wall outlet which is female, one end of the cord should be fitted with a male plug. The other end of the cord should have a female receptacle. The input connector of the power supply should have a male receptacle to fit the female receptacle of the cord. The power-output connector on the power supply should be a female socket. A male plug to fit this socket should be connected to the cable going to the equipment. The opposite end of the cable should be fitted with a female connector, and the series should terminate with a male connector on the equipment. If connections are made in this manner, there should be no "live" exposed contacts at any point, regardless of where a disconnection may be made.

Rectifier filament leads should be kept short

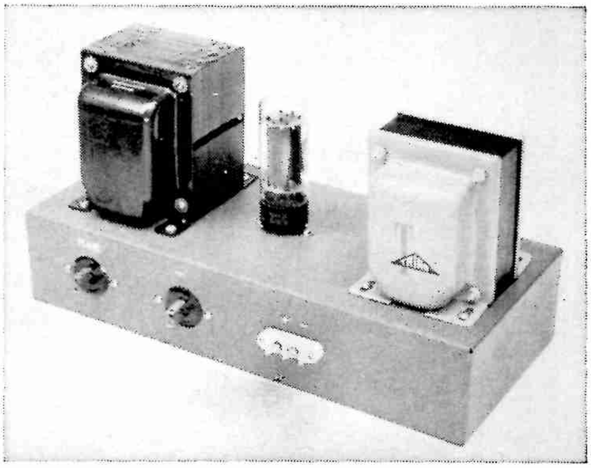


Fig. 7-33—A typical low-voltage power supply. The two a.c. connectors permit independent control of filament and high voltage.

to assure proper voltage at the rectifier socket, through a metal chassis, grommet-lined clearance holes will serve for voltages up to 500 or 750, but ceramic feed-through insulators should be used for higher voltages. Bleeder and voltage-dropping resistors should be placed where they are open to air circulation. Placing them in confined space reduces the rating.

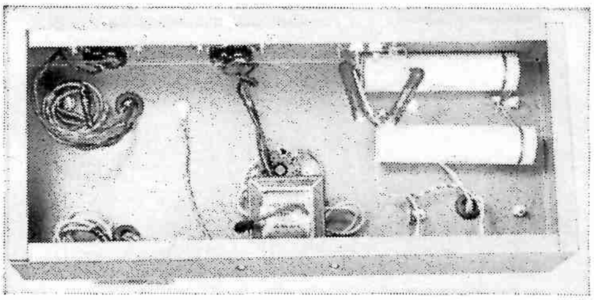


Fig. 7-34. A bottom view of the low-voltage power supply. The separate filament transformer is mounted against the lower wall of the chassis. The electrolytic filter capacitors are mounted on terminal strips. Rubber grommets are used where wires pass through the chassis.

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. When using a combination power transformer, high voltage may be turned off without turning the filaments off by using a switch between the transformer center tap and chassis. This switch should be of the rotary type with good insulation between contacts. The shaft of the switch *must* be grounded.

● SAFETY PRECAUTIONS

All power supplies in an installation should be fed through a single main power-line switch so that all power may be cut off quickly, either before working on the equipment, or in case of an accident. Spring-operated switches or relays are not sufficiently reliable for this important service. Foolproof devices for cutting off all power to the transmitter and other equipment are shown in Fig. 7-37. The arrangements shown in Fig. 7-37A and B are similar circuits for two-wire (115-volt) and three-wire (230-volt) systems. *S* is an enclosed double-throw knife switch of the sort usually used as the entrance switch in house installations. *J* is a standard a.c. outlet and *P* a shorted plug to fit the outlet. The switch should be located prominently in plain sight and mem-

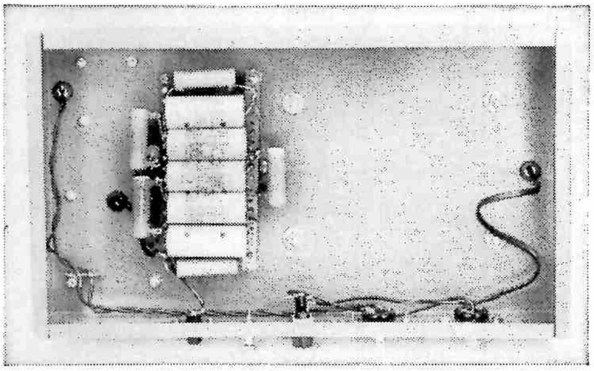


Fig. 7-36—Bottom view of the high-voltage supply. The electrolytic capacitors (connected in series) are mounted on an insulating board. Voltage-equalizing resistors are connected across each capacitor. Separate input connectors are provided for filament and plate power.

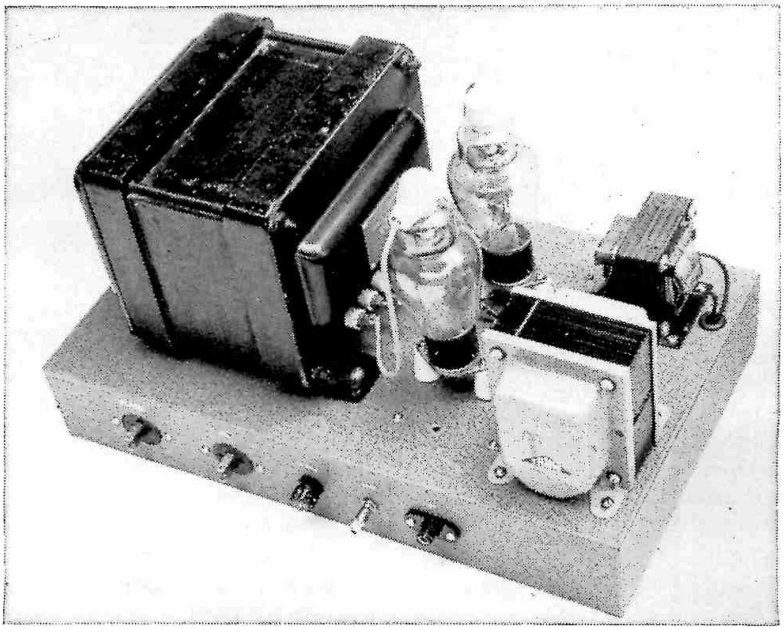


Fig. 7-35—A typical high-voltage supply. The sockets for the 866A mercury-vapor rectifier tubes are spaced from the metal chassis by small cone insulators. Note the insulated tube plate connectors, the safety high-voltage output terminal and the fuse.

Power-Supply Construction

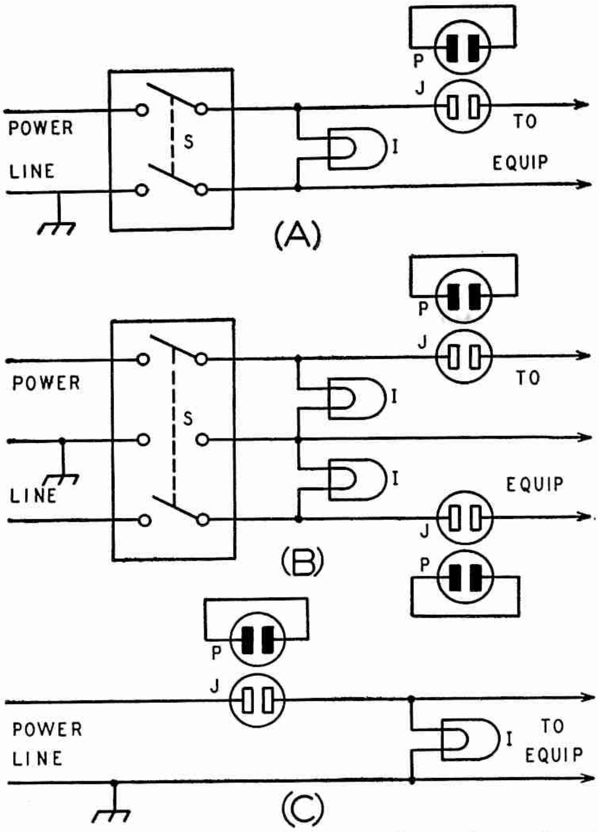


Fig. 7-37—Reliable arrangements for cutting off all power to the transmitter. S is an enclosed double-pole knife-type switch, J a standard a.c. outlet. P a shorted plug to fit the outlet and I a red lamp.

A is for a two-wire 115-volt line, B for a three-wire 230-volt system, and C a simplified arrangement for low-power stations.

bers of the household should be instructed in its location and use. *I* is a red lamp located alongside the switch. Its purpose is not so much to serve as a warning that the power is on as it is to help in identifying and quickly locating the switch should it become necessary for someone else to cut the power off in an emergency.

The outlet *J* should be placed in some corner out of sight where it will not be a temptation for children or others to play with. The shorting plug can be removed to open the power circuit if there are others around who might inadvertently throw the switch while the operator is working on the rig. If the operator takes the plug with him, it will prevent someone from turning on the power in his absence and either injuring themselves or the equipment or perhaps starting a fire. Of utmost importance is the fact that the outlet *J must* be placed in the *ungrounded* side of the line.

Those who are operating low power and feel that the expense or complication of the switch isn't warranted can use the shorted-plug idea as the main power switch. In this case, the outlet should be located prominently and identified by a signal light, as shown in Fig. 7-37C.

The test bench ought to be fed through the main power switch, or a similar arrangement at the bench, if the bench is located remote from the transmitter.

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 capacitors will be discharged when the high-voltage transformer is turned off.

Selenium-Rectifier Table

All types listed below are rated as follows: Max. input r.m.s. volts — 130, Max. peak inverse volts — 380. Series resistors of 47 ohms are recommended for units rated at less than 65 ma., 22 ohms for 75- and 100-ma. units, 15 ohms for 150-ma. units, and 5 ohms for all higher-current units.

D.C. Ma. Output	*Manufacturer*					
	A	*B*	*C*	*D*	*E*	*F*
20	1159	....	8S20	....	...	....
30	....	....	....	8Y1	...	....
35	....	....	8S35	....	...	....
50	....	RS65Q	....	....	50	....
65	1002A	RS65	6S65	8J1	65	NA-5
75	1003A	RS75	6S75	5M4	75	NB-5
100	1004A	RS100	6S100	5M1	100	NC-5
150	1005A	RS150	6S150	5P1	150	ND-5
200	1006A	RS200	6S200	5R1	200	NE-5
250	1028A	RS250	6S250	5Q1	250	NF-5
300	1090A	RS300	6S300	6Q4	300	....
350	1023	RS350	6S350	5QS1	...	NK-5
400	1130	RS400	6S400	5S2	400	NH-5
450	....	RS450	6S450	....	...	NJ-5
500	1179	RS500	6S500	5S1	500	....
600	....	....	....	....	600	....
1000	....	RS1000	....	....	...	....

A — Federal. B — International. C — Mallory. D — Radio Receptor. E — Sarkes-Tarzian. F — Sylvania.

Silicon Rectifier Table

JETEC Type	*Max. R.M.S. Input Volts*	*Max. D.C. Load Current*
1N1082	140	500 ma.
1N1084	280	500 ma.
1N1109	840	425 ma.
1N1110	1120	400 ma.
1N1113	1960	325 ma.
*M150	130	150 ma.
*M500	130	500 ma.

* Sarkes-Tarzian type number.

Germanium Rectifier Table

(All 300 ma. d.c. output)

JETEC Type	*Max. R.M.S. Volts Input*	*JETEC Type*	*Max. R.M.S. Volts Input*	*JETEC Type*	*Max. R.M.S. Volts Input*
1N600	70	1N600A	70	1N611	210
1N601	105	1N603A	210	1N613	350
1N602	140	1N604A	280	1N607A	35
1N604	280	1N605A	350	1N608A	70
1N605	350	1N607	35	1N611A	210
1N599A	35	1N608	70	1N612A	280

CHAPTER 8

Keying and Break-In

Section 12.133 of the FCC regulations says ". . . The frequency of the emitted . . . wave shall be as constant as the state of the art permits." It also says ". . . spurious radiation shall not be of sufficient intensity to cause interference in receiving equipment of good engineering design including adequate selectivity characteristics, which is tuned to a frequency or frequencies outside the frequency band of emission normally required for the type of emission being employed by the amateur station."

The state of the art is such that an emitted wave can be mighty stable, yet many code (and phone) stations show f.m. and chirp that leaves them open to a citation by the Commission. Key clicks (and splatter) represent violations of the spurious radiation clause, and it isn't hard to find evidences of them in any of the ham bands.

There are four factors that have to be considered in the keying of a transmitter. They are r.f. clicks, envelope shape, chirp and backwave.

R.F. Clicks

When any circuit carrying d.c. or a.c. is closed or broken, the small or large spark (depending upon the voltage and current) generates r.f. during the instant of make or break. This r.f. covers a frequency range of many megacycles. When a transmitter is keyed, the spark at the key (and relay, if used) causes a click in the receiver. *This click has no effect on the transmitted signal.* Since it occurs at the same time that a click (if any) appears on the transmitter output, it must be removed if one is to listen critically to his own signal within the shack. A small r.f. filter is required at the contacts of the key (and relay); typical circuits and values are shown in Fig. 8-1. To check the effectiveness of the r.f. filter, listen on a lower-frequency band than the transmitter is tuned to, with a short antenna and the gain backed off.

Envelope Shape

The key clicks that go out on the air with the signal are controlled by the shape of the envelope of the signal. The envelope is the outline of the oscilloscope pattern of your transmitter output, but an oscilloscope isn't needed to observe the effects. Fig. 8-2 shows representative scope patterns that might be obtained with a given transmitter under various conditions.

One should understand that the *on-the-air* clicks are determined by the shaping, while the r.f. clicks caused by the spark at the key can only be heard in the station receiver and possibly a broadcast receiver in the same house or apartment.

Chirp

The frequency-stability reference in the opening paragraph refers to the "chirp" observed on many signals. This is caused by a change in frequency of the signal during a single dot or dash. Chirp is an easy thing to detect if you know how to listen for it, although it is amazing how some operators will listen to a signal and say it has no chirp when it actually has. The easiest way to detect chirp is to tune in the code signal at a low beat note and listen for any change in frequency during a dash. The lower the beat note, the easier it is to detect the frequency change. Listening to a harmonic of the signal will accentuate the frequency change.

The main reason for minimizing chirp, aside from complying with the letter of the regulations, is one of pride, since a properly shaped chirp-free signal is a pleasure to copy and is likely to attract attention by its rarity. Chirps cannot be observed on an oscilloscope pattern of the envelope.

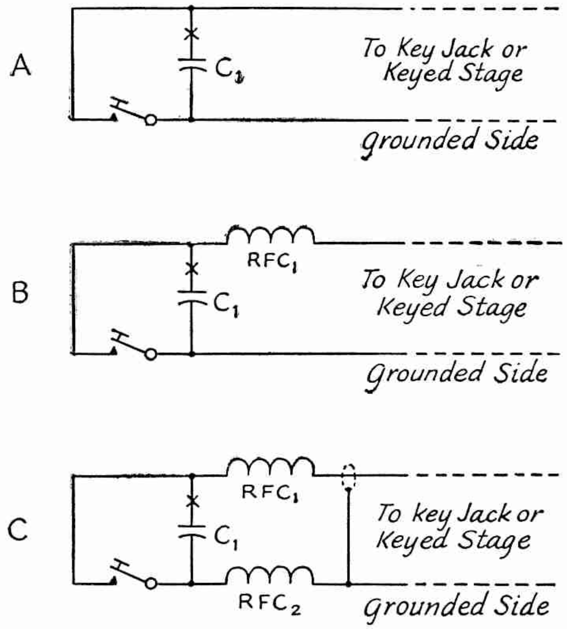


Fig. 8-1—Typical filter circuits to apply at the key (and relay, if used) to minimize r.f. clicks. The simplest circuit (A) is a small capacitor mounted at the key. If this proves insufficient, an r.f. choke can be added to the ungrounded lead (B) or in both leads (C). The value of C_1 is .001 to .01 µf., RFC_1 and RFC_2 can be 0.5 to 2.5 mh., with a current-carrying ability sufficient for the current in the keyed circuit. In difficult cases another small capacitor may be required on the other side of the r.f. choke or chokes. In all cases the r.f. filter should be mounted right at the key or relay terminals; sometimes the filter can be concealed under the key. When cathode or center-tap keying is used, the resistance of the r.f. choke or chokes will add cathode bias to the keyed stage, and in this case a high-current low-resistance choke may be required, or compensating reduction of the grid-leak bias (if it is used) may be needed.

A visible spark on "make" can often be reduced by the addition of a small (10 to 100 ohms) resistor in series with C_1 (inserted at point "x"). Too high a value of resistor reduces the arc-suppressing effect on "break."